**HANDBOOK**
**FOR**
**HEADEND PRODUCTS**
**TRAINING SCHOOL**

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## FORWARD

The contents of this handbook contain reprints of several papers and application notes along with the basic material in each section. To avoid massive renumbering of figures and tables, all basic section figures and table numbers are prefixed with the section number followed by a dash, whereas figure and table numbers for papers and application notes retain their original numbering scheme. Thus, if a figure or table is referenced that does not include the section number followed by a dash, you will know that this is a reference within that paper or application note. However, all pages (including papers and application notes) are paginated by section.

# SECTION 1
# BASIC CATV
# SYSTEM CONCEPTS

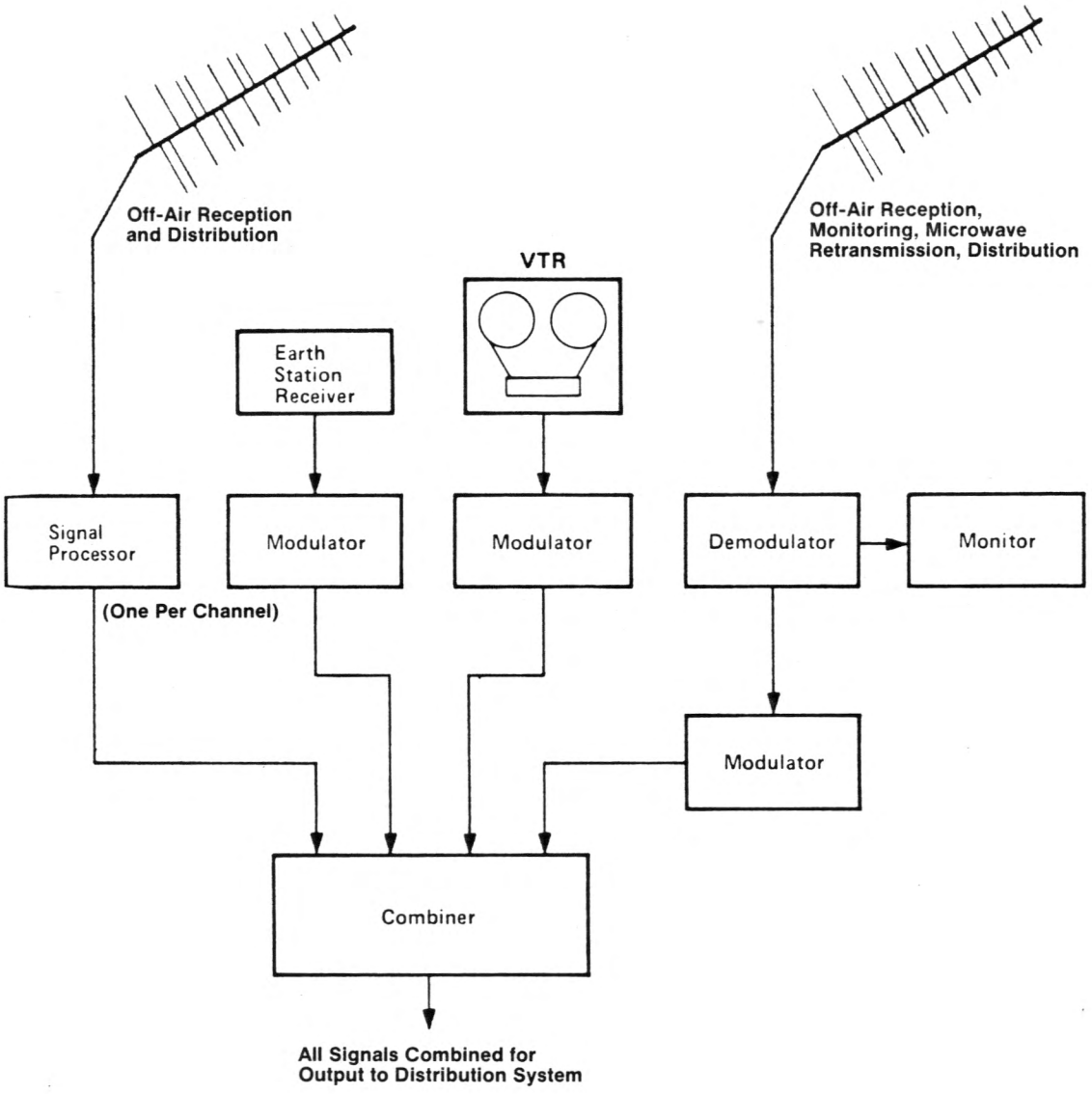



**Figure 1-1. Headend Block Diagram**

# SECTION 2
# OFF-AIR ANTENNAS AND PREAMPLIFIERS

## Equipment Configurations

As long as two stations of interest are within the reception pattern (beamwidth), the two signal levels are acceptable, and the antenna exhibits sufficient bandwidth to receive both channels, one antenna can be used to furnish the headend with the two channels. In the illustration below Antenna 1 is aimed between Stations A and B to receive signals from both A and B.

The bearing to Station C is too far removed from the other two stations to use Antenna 1, even if its bandwidth were sufficient. Therefore, a second antenna would be required to receive the signal from C.

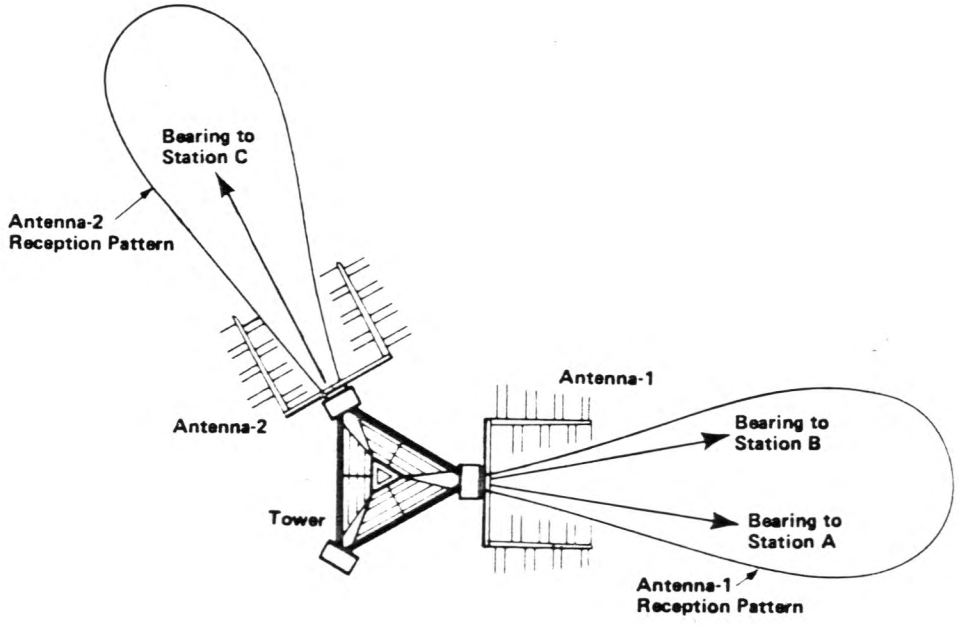



**Figure 2-1. Tower Orientation of Off-Air Antennas**

If the signals from A and B were of sufficient level that the headend signal-processing equipment could adequately produce a high quality television signal for distribution, no amplification of the received signal would be necessary. Scientific-Atlanta recommends that a preamplifier not be used just to "boost" the signal if a usable, quality signal can be obtained without amplification.

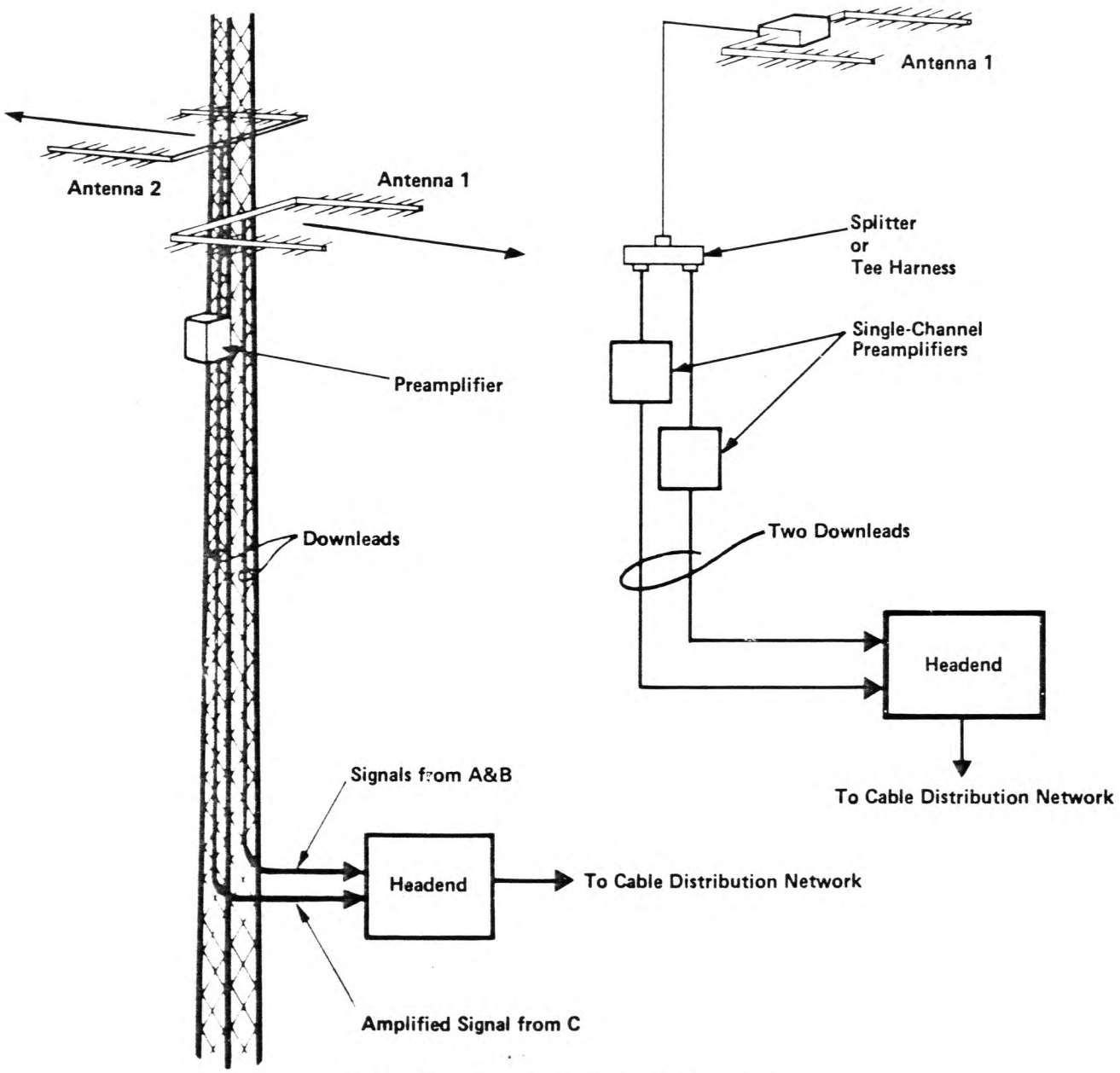



**Figure 2-2. Tower-Mounted Amplification**

However, if the signal level were marginal, then a tower-mounted, single-channel preamplifier, such as that shown below connected to Antenna 2, should be included to compensate for the downlead cable loss. This preamplifier will selectively amplify only one channel, while effectively rejecting others.

Another situation might require preamplification of both Signals A and B from Antenna 1. In this case, a signal splitter would be required between the antenna and two preamplifiers, one for each signal. Two downlead cables feeding two signal processors in the headend would be required.

## Computer-Aided Signal Survey

Extremely costly errors can be eliminated by knowing the quality of the signals available at your tower site before construction is started. Scientific-Atlanta offers a unique service to the CATV operator through our Computer-Aided Signal Survey, CASS takes your tower location and height and provides you with a comprehensive listing of all TV stations within a 200-mile radius of the proposed tower site. This listing contains such pertinent information as channel number, call letters, transmitter location, network affiliation, effective radiated power, transmitting antenna height above average terrain, bearing to the transmitter, and airline distance between this transmitter and your tower. Of prime importance, CASS also predicts the average received signal level, level of co-channel interference to be expected, and even recommends the correct antenna and preamplifier to use for optimum off-air performance.

Station information on all known commercial and educational TV stations and construction permits in the United States and Canada within about 200 miles of your tower location is included in this survey. This information is updated weekly using the latest Television Digest Weekly TV Addenda. All signal levels are calculated from the top of your tower so that antennas for the weakest signals can be analyzed at maximum height.

CASS is based on propagation equations developed by the National Bureau of Standards, and published in their Technical Note 101, "Transmission Loss Predictions for Tropospheric Communications Circuits." It is the most accurate theoretical survey available. Nevertheless, there are numerous factors that may affect signal propagation and picture quality at any given location. For this reason, there may be differences between the signal level predicted and that which is actually received. In normal "smooth earth" locations these variations are generally within 6 dB and small enough to be overlooked. But there is always the possibility of significant deviation. For that reason, an actual on-site survey is strongly recommended before a tower site is selected.

So, to arrange for your CASS, please contact the CATV experts at Scientific-Atlanta. In planning your off-air equipment, you really can't afford not to.

SCIENTIFIC-ATLANTA, INC.
P.O. Box 105027
Atlanta, GA. 30324
Cable Communications
Marketing Dept.

REQUEST FOR COMPUTER SIGNAL SURVEY

DATE: _____

By: _____

Co-ordinates _____ North Latitude _____ West Longitude

Company _____

SIGNAL SURVEY SITE _____
City & State

Tower Height _____
(Feet above average terrain)

MAIL SURVEY TO:

Name _____

Company _____

Address _____ Zip _____

City _____ State _____

## SYMBOLS AND ABBREVIATIONS

| | |
|---|---|
| CH | VHF or UHF Station Channel Number |
| + | Positive Carrier Offset (+10 kHz) |
| | Zero Carrier Offset (No Symbol) |
| – | Negative Carrier Offset (-10 kHz) |
| CALL | Station Call Letters |
| CITY | Station Location |
| NET | Network Affiliation |
| ABC | American Broadcasting Company |
| CBS | Columbia Broadcasting System |
| NBC | National Broadcasting Company |
| A-C | ABC-CBS |
| A-N | ABC-NBC |
| C-N | CBS-NBC |
| ACN | ABC-CBS-NBC |
| ETV | Educational |
| IND | Independent |
| CBC | Canadian Broadcasting Corporation |
| CTV | CTV Television Network |
| CP | Construction Permit (Not on Air) |
| RAD KW | Station Radiated Power (Horizontal Visual) in Kilowatts |
| HT | Station Antenna Height Above Average Terrain (Feet) |
| BNG | True Bearing to TV Station (Degrees) |
| DIST | Distance to TV Station (Statute Miles) |
| ARL (DB) | Calculated Average Received Signal Level (dBmV) |
| ARL (UV) | Calculated Average Received Signal Level (Microvolts) |

## IMPORTANT INFORMATION

1. In each channel listing, the first station listed is nearest, and is the desired station; subsequent stations are listed in order of increasing distance from the tower site. The antenna is "pointed" at the desired station.
2. Received signal levels take into account the gain and directivity of the antenna.
3. The word "RECOMMENDED" does not appear when received signal levels are below acceptable minimums for carriage on a cable TV system.
4. UHF stations between 150 and 300 miles from the CATV tower are listed without signal level calculations.
5. A new "RECOMMENDED ANTENNA" heading is printed whenever the antenna type changes. Desired stations listed under one "RECOMMENDED ANTENNA" heading may require separate antennas. Two stations with high received signal levels and bearings within 10° for a "QCS," or 40° for a "QCA" can usually be received using one antenna.
6. The survey assumes "smooth earth" conditions and therefore, is usually somewhat inaccurate in rugged, mountainous terrain.
7. Antenna height is always equal to the maximum tower height in feet above average terrain elevation. As a general "rule-of-thumb", signal level increases 6 dB each time the height is doubled.
8. The survey does not test for adjacent channel interference. Possible adjacent channel interference must be considered whenever an adjacent channel under the same "ANTENNA" heading as the desired station has a received signal level greater than the desired station.
9. The summary sheet lists only those stations with predicted signal levels $> -6$ dBmV. When preamplifiers are recommended (in brackets), the indicated signal level includes the preamplifier gain.
10. The summary sheet indicates the predicted severity of co-channel interference:

| | |
|---|---|
| "CONTINUOUS" | $>$ 40 dB difference in levels |
| "FREQUENT" | 41-50 dB difference in levels |
| "OCCASIONAL" | 51-60 dB difference in levels |
| NONE ANTICIPATED | $>$ 60 dB difference in levels |

THIS REPORT PREPARED SPECIFICALLY FOR

SCIENTIFIC ATLANTA INC.

COMPUTER SIGNAL SURVEY
3/ 5/81

HEADEND LOCATION: NORCROSS GEORGIA TEST SITE

GEOGRAPHICAL CO-ORDINATES: 33 56 0 NORTH LATITUDE
84 12 0 WEST LONGITUDE

RECEIVE ANTENNA HEIGHT: 100 FEET

SCIENTIFIC-ATLANTA, INC.
CABLE COMMUNICATIONS DIVISION
P. O. BOX 105027
ATLANTA, GEORGIA 30348
TEL. (404) 449-2000

SIGNAL SURVEY COURTESY OF SCIENTIFIC-ATLANTA, INC. PAGE 1

RECOMMENDED ANTENNA: QCA-2

| CH | CALL | CITY | NET | RAD KW | HT | BNG | DIST | ARL(UV) | (DB) |
|---|---|---|---|---|---|---|---|---|---|
| 2 | WSB | ATLANTA GA | NBC | 100.0 | 1037 | 218 | 14.9 | 84644.1 | 38.6 |
| 2+ | WSJK | SNEEDVILLE TN | ETV | 100.0 | 1763 | 19 | 178.6 | 1.3 | -57.6 |

ANTENNA: QCS-2

| CH | CALL | CITY | NET | RAD KW | HT | BNG | DIST | ARL(UV) | (DB) |
|---|---|---|---|---|---|---|---|---|---|
| 3 | WRCB | CHATTANOOGA TN | NBC | 100.0 | 1070 | 324 | 105.8 | 148.6 | -16.6 |
| 3 | WRBL | COLUMBUS GA | CBS | 100.0 | 1790 | 197 | 116.1 | 10.6 | -39.5 |
| 3 | WBTV | CHARLOTTE NC | CBS | 100.0 | 1090 | 61 | 199.0 | 1.6 | -56.1 |

ANTENNA: QCS-4

| CH | CALL | CITY | NET | RAD KW | HT | BNG | DIST | ARL(UV) | (DB) |
|---|---|---|---|---|---|---|---|---|---|
| 4- | WFBC | GREENVILLE SC | NBC | 100.0 | 2000 | 48 | 121.9 | 140.3 | -17.1 |

RECOMMENDED ANTENNA: QCA-4

| CH | CALL | CITY | NET | RAD KW | HT | BNG | DIST | ARL(UV) | (DB) |
|---|---|---|---|---|---|---|---|---|---|
| 5- | WAGA | ATLANTA GA | CBS | 100.0 | 1070 | 219 | 12.1 | 129051.9 | 42.2 |

ANTENNA: QCS-4

| CH | CALL | CITY | NET | RAD KW | HT | BNG | DIST | ARL(UV) | (DB) |
|---|---|---|---|---|---|---|---|---|---|
| 6+ | WJBF | AUGUSTA GA | ABC | 100.0 | 1375 | 104 | 140.5 | 50.2 | -26.0 |
| 6 | WATE | KNOXVILLE TN | NBC | 100.0 | 1490 | 6 | 143.7 | 2.7 | -51.2 |
| 6- | WBRC | BIRMINGHAM AL | ABC | 100.0 | 1120 | 259 | 152.4 | 2.2 | -53.2 |

RECOMMENDED ANTENNA: QCS-7

| CH | CALL | CITY | NET | RAD KW | HT | BNG | DIST | ARL(UV) | (DB) |
|---|---|---|---|---|---|---|---|---|---|
| 7- | WCIQ | MOUNT CHEAHA AL | ETV | 316.0 | 2000 | 252 | 97.5 | 743.8 | -2.6 |
| 7+ | WSPA | GREENVILLE SC | CBS | 316.0 | 2000 | 51 | 138.1 | 3.8 | -48.4 |

RECOMMENDED ANTENNA: QCA-7

| CH | CALL | CITY | NET | RAD KW | HT | BNG | DIST | ARL(UV) | (DB) |
|---|---|---|---|---|---|---|---|---|---|
| 8+ | WGTV | ATHENS GA | ETV | 288.0 | 1080 | 160 | 9.4 | 151437.4 | 43.6 |
| 8 | WSLA | SELMA AL | CBS | 53.7 | 360 | 238 | 197.6 | 0.2 | -74.6 |

ANTENNA: QCS-7

| CH | CALL | CITY | NET | RAD KW | HT | BNG | DIST | ARL(UV) | (DB) |
|---|---|---|---|---|---|---|---|---|---|
| 9 | WTVC | CHATTANOOGA TN | ABC | 316.0 | 1040 | 323 | 106.0 | 97.8 | -20.2 |
| 9+ | WTVM | COLUMBUS GA | ABC | 284.0 | 1650 | 197 | 116.1 | 5.2 | -45.7 |
| 9- | WVAN | SAVANNAH GA | ETV | 316.0 | 1050 | 129 | 193.8 | 1.0 | -59.6 |

ANTENNA: QCS-7

| CH | CALL | CITY | NET | RAD KW | HT | BNG | DIST | ARL(UV) | (DB) |
|---|---|---|---|---|---|---|---|---|---|
| 10+ | WBIR | KNOXVILLE TN | CBS | 316.0 | 1790 | 6 | 143.8 | 50.7 | -25.9 |
| 10- | WBIQ | BIRMINGHAM AL | ETV | 316.0 | 1050 | 259 | 152.4 | 1.9 | -54.2 |
| 10 | WALB | ALBANY GA | NBC | 316.0 | 964 | 174 | 180.8 | 1.2 | -58.5 |
| 10- | WIS | COLUMBIA SC | NBC | 316.0 | 1546 | 85 | 197.5 | 1.1 | -59.5 |

RECOMMENDED ANTENNA: QCA-7

| CH | CALL | CITY | NET | RAD KW | HT | BNG | DIST | ARL(UV) | (DB) |
|---|---|---|---|---|---|---|---|---|---|
| 11 | WXIA | ATLANTA GA | ABC | 316.0 | 1050 | 212 | 14.4 | 94600.1 | 39.5 |

ANTENNA: QCS-7

| CH | CALL | CITY | NET | RAD KW | HT | BNG | DIST | ARL(UV) | (DB) |
|---|---|---|---|---|---|---|---|---|---|
| 12- | WDEF | CHATTANOOGA TN | CBS | 316.0 | 1260 | 323 | 104.7 | 103.3 | -19.7 |
| 12+ | WRDW | AUGUSTA GA | CBS | 316.0 | 1590 | 104 | 140.2 | 2.7 | -51.3 |
| 12 | WSFA | MONTGOMERY AL | NBC | 316.0 | 2000 | 221 | 176.6 | 1.5 | -56.5 |

ANTENNA: QCS-7

| CH | CALL | CITY | NET | RAD KW | HT | BNG | DIST | ARL(UV) | (DB) |
|---|---|---|---|---|---|---|---|---|---|
| 13+ | WMAZ | MACON GA | A-C | 316.0 | 785 | 155 | 89.5 | 119.2 | -18.5 |
| 13- | WLOS | GREENVILLE SC | ABC | 178.0 | 2804 | 39 | 129.9 | 3.7 | -48.6 |
| 13- | WVTM | BIRMINGHAM AL | NBC | 316.0 | 1360 | 259 | 152.2 | 1.9 | -54.3 |

| CH | CALL | CITY | NET | RAD KW | HT | BNG | DIST | ARL(UV) | (DB) |
|---|---|---|---|---|---|---|---|---|---|
| 14 | WEBA | ALLENDALE SC | ETV | 557.0 | 800 | 107 | 169.2 | | |
| 14 | WABW | PELHAM GA | ETV | 363.0 | 1242 | 178 | 193.3 | | |

ANTENNA: UCA-8

| CH | CALL | CITY | NET | RAD KW | HT | BNG | DIST | ARL(UV) | (DB) |
|---|---|---|---|---|---|---|---|---|---|
| 15 | WDCO | COCHRAN GA | ETV | 363.0 | 1149 | 151 | 114.9 | 37.3 | -28.6 |

ANTENNA: UCA-8

| CH | CALL | CITY | NET | RAD KW | HT | BNG | DIST | ARL(UV) | (DB) |
|---|---|---|---|---|---|---|---|---|---|
| 16 | WGGS | GREENVILLE SC | IND | 525.0 | 1149 | 55 | 123.3 | 35.1 | -29.1 |

RECOMMENDED ANTENNA: QCAUHF

| CH | CALL | CITY | NET | RAD KW | HT | BNG | DIST | ARL(UV) | (DB) |
|---|---|---|---|---|---|---|---|---|---|
| 17 | WTBS | ATLANTA GA | IND | 1510.0 | 1093 | 226 | 15.0 | 90290.7 | 39.1 |
| 17 | WUNE | LINVILLE NC | ETV | 38.0 | 1540 | 42 | 198.5 | | |
| 18 | WCLP | CHATSWORTH GA | ETV | 447.0 | 1854 | 333 | 63.7 | 11451.6 | 21.2 |
| 18 | WDHN | DOTHAN AL | ABC | 398.0 | 730 | 200 | 196.8 | | |

ANTENNA: UCA-8

| CH | CALL | CITY | NET | RAD KW | HT | BNG | DIST | ARL(UV) | (DB) |
|---|---|---|---|---|---|---|---|---|---|
| 19 | WHNT | HUNTSVILLE AL | CBS | 1002.0 | 1750 | 293 | 144.1 | 35.8 | -28.9 |
| 19 | WLTX | COLUMBIA SC | CBS | 1046.0 | 637 | 87 | 183.5 | | |

ANTENNA: UCA-8

| CH | CALL | CITY | NET | RAD KW | HT | BNG | DIST | ARL(UV) | (DB) |
|---|---|---|---|---|---|---|---|---|---|
| 20 | WCES | WRENS GA | ETV | 331.0 | 1483 | 112 | 119.5 | 35.9 | -28.9 |
| 20 | WCOV | MONTGOMERY AL | CBS | 550.0 | 740 | 228 | 163.6 | | |

| CH | CALL | CITY | NET | RAD KW | HT | BNG | DIST | ARL(UV) | (DB) |
|---|---|---|---|---|---|---|---|---|---|
| 22 | WCTE | COOKEVILLE TN | ETV | 427.0 | 1390 | 338 | 167.7 | | |

ANTENNA: UCA-8

| CH | CALL | CITY | NET | RAD KW | HT | BNG | DIST | ARL(UV) | (DB) |
|---|---|---|---|---|---|---|---|---|---|
| 25 | WACS | DAWSON GA | ETV | 363.0 | 1078 | 189 | 139.3 | 18.4 | -34.7 |
| 25 | WHIQ | HUNTSVILLE AL | ETV | 631.0 | 1173 | 293 | 144.2 | 1.3 | -58.0 |
| 25 | WOLO | COLUMBIA SC | ABC | 2344.0 | 688 | 86 | 184.5 | | |

ANTENNA: UCA-8

| CH | CALL | CITY | NET | RAD KW | HT | BNG | DIST | ARL(UV) | (DB) |
|---|---|---|---|---|---|---|---|---|---|
| 26 | WATU | AUGUSTA GA | NBC | 933.0 | 1590 | 104 | 140.2 | 34.9 | -29.1 |
| 26 | WTVK | KNOXVILLE TN | ABC | 1620.0 | 1270 | 6 | 144.2 | 2.1 | -53.6 |
| 26 | WAIQ | MONTGOMERY AL | ETV | 229.0 | 600 | 229 | 161.6 | | |

RECOMMENDED ANTENNA: UCA-8

| CH | CALL | CITY | NET | RAD KW | HT | BNG | DIST | ARL(UV) | (DB) |
|---|---|---|---|---|---|---|---|---|---|
| 28 | WJSP | COLUMBUS GA | ETV | 921.0 | 1510 | 201 | 80.0 | 1682.9 | 4.5 |
| 29 | WNTV | GREENVILLE SC | ETV | 339.0 | 1142 | 55 | 123.4 | 26.7 | -31.5 |

RECOMMENDED ANTENNA: QCAUHF

| CH | CALL | CITY | NET | RAD KW | HT | BNG | DIST | ARL(UV) | (DB) |
|---|---|---|---|---|---|---|---|---|---|
| 30 | WETV | ATLANTA GA | ETV | 776.0 | 1090 | 223 | 11.9 | 70784.4 | 37.0 |
| 30 | WNSC | ROCK HILL SC | ETV | 575.0 | 691 | 70 | 191.8 | | |

ANTENNA: UCA-8

| CH | CALL | CITY | NET | RAD KW | HT | BNG | DIST | ARL(UV) | (DB) |
|---|---|---|---|---|---|---|---|---|---|
| 31 | WAAY | HUNTSVILLE AL | ABC | 674.5 | 1792 | 293 | 144.2 | 28.3 | -31.0 |

| CH | CALL | CITY | NET | RAD KW | HT | BNG | DIST | ARL(UV) | (DB) |
|---|---|---|---|---|---|---|---|---|---|
| 32 | WKAB | MONTGOMERY AL | ABC | 813.0 | 830 | 228 | 160.9 | | |

ANTENNA: UCA-8

| CH | CALL | CITY | NET | RAD KW | HT | BNG | DIST | ARL(UV) | (DB) |
|---|---|---|---|---|---|---|---|---|---|
| 33 | WUNF | ASHEVILLE NC | ETV | 3.6 | 2620 | 38 | 131.6 | 4.0 | -47.9 |
| 33 | WCFT | TUSCALOOSA AL | CBS | 1000.0 | 540 | 255 | 196.1 | | |

| CH | CALL | CITY | NET | RAD KW | HT | BNG | DIST | ARL(UV) | (DB) |
|---|---|---|---|---|---|---|---|---|---|
| 35 | WRLK | COLUMBIA SC | ETV | 513.0 | 1030 | 85 | 187.2 | | |

RECOMMENDED ANTENNA: QCAUHF

| CH | CALL | CITY | NET | RAD KW | HT | BNG | DIST | ARL(UV) | (DB) |
|---|---|---|---|---|---|---|---|---|---|
| 36 | WATL | ATLANTA GA | IND | 2030.0 | 1086 | 223 | 11.9 | 107654.4 | 40.6 |

ANTENNA: UCA-8

| CH | CALL | CITY | NET | RAD KW | HT | BNG | DIST | ARL(UV) | (DB) |
|---|---|---|---|---|---|---|---|---|---|
| 38 | WYEA | COLUMBUS GA | NBC | 1070.0 | 1310 | 201 | 109.3 | 77.7 | -22.2 |

ANTENNA: UCA-8

| CH | CALL | CITY | NET | RAD KW | HT | BNG | DIST | ARL(UV) | (DB) |
|---|---|---|---|---|---|---|---|---|---|
| 40 | WHMA | ANNISTON AL | CBS | 464.0 | 880 | 260 | 94.3 | 66.2 | -23.6 |
| 40 | WAIM | ANDERSON SC | ABC | 1620.0 | 840 | 65 | 102.9 | 4.9 | -46.1 |

ANTENNA: UCA-8

| CH | CALL | CITY | NET | RAD KW | HT | BNG | DIST | ARL(UV) | (DB) |
|---|---|---|---|---|---|---|---|---|---|
| 41 | WCWB | MACON GA | NBC | 760.0 | 787 | 156 | 89.4 | 95.3 | -20.4 |

| CH | CALL | CITY | NET | RAD KW | HT | BNG | DIST | ARL(UV) | (DB) |
|---|---|---|---|---|---|---|---|---|---|
| 42 | WBMG | BIRMINGHAM AL | CBS | 891.0 | 1151 | 259 | 152.9 | | |

| CH | CALL | CITY | NET | RAD KW | HT | BNG | DIST | ARL(UV) | (DB) |
|---|---|---|---|---|---|---|---|---|---|
| 43 | WGIQ | LOUISVILLE AL | ETV | 708.0 | 940 | 205 | 167.8 | | |

ANTENNA: UCA-8

| CH | CALL | CITY | NET | RAD KW | HT | BNG | DIST | ARL(UV) | (DB) |
|---|---|---|---|---|---|---|---|---|---|
| 45 | WTCI | CHATTANOOGA TN | ETV | 617.0 | 950 | 325 | 107.3 | 49.6 | -26.1 |

RECOMMENDED ANTENNA: QCAUHF

| CH | CALL | CITY | NET | RAD KW | HT | BNG | DIST | ARL(UV) | (DB) |
|---|---|---|---|---|---|---|---|---|---|
| 46 | WANX | ATLANTA GA | IND | 1290.0 | 1070 | 223 | 11.9 | 78054.7 | 37.8 |

ANTENNA: UCA-8

| CH | CALL | CITY | NET | RAD KW | HT | BNG | DIST | ARL(UV) | (DB) |
|---|---|---|---|---|---|---|---|---|---|
| 48 | WAFF | HUNTSVILLE AL | NBC | 525.0 | 1900 | 293 | 143.6 | 24.7 | -32.1 |

| CH | CALL | CITY | NET | RAD KW | HT | BNG | DIST | ARL(UV) | (DB) |
|---|---|---|---|---|---|---|---|---|---|
| 49+ | WRET | SPARTANBURG SC | ETV | 832.0 | 970 | 63 | 150.6 | | |

ANTENNA: UCA-8

| CH | CALL | CITY | NET | RAD KW | HT | BNG | DIST | ARL(UV) | (DB) |
|---|---|---|---|---|---|---|---|---|---|
| 55 | WCPT | CROSSVILLE TN | CBC | 18.6 | 1180 | 345 | 139.9 | 3.8 | -48.4 |

ANTENNA: UCA-8

| CH | CALL | CITY | NET | RAD KW | HT | BNG | DIST | ARL(UV) | (DB) |
|---|---|---|---|---|---|---|---|---|---|
| 61 | WRIP | CHATTANOOGA TN | IND | 457.0 | 980 | 326 | 107.3 | 41.5 | -27.6 |

SCIENTIFIC ATLANTA INC. 100 FT, 69 STATIONS.

```
.............. SUMMARY REPORT ..............

NORCROSS GEORGIA TEST SITE                    PAGE 1

    33 56  0  NN        84 12  0   W
       100 FEET      3/ 5/81

CH  CALL   CITY           NET DEG  MILES   DBMV

QCA-2
 2   WSB ATLANTA GA       NBC 218   14.9   38.6

QCA-4
 5- WAGA ATLANTA GA       CBS 219   12.1   42.2

QCS-7
 7- WCIQ MOUNT CHEAHA AL ETV 252   97.5   -2.6          PREAMP
       FREQUENT CO-CHANNEL FROM: WSPA GREENVILLE SC     45.8DB  201DEG

QCA-7
 8+ WGTV ATHENS GA        ETV 160    9.4   43.6
11  WXIA ATLANTA GA       ABC 212   14.4   39.5

QCAUHF
17  WTBS ATLANTA GA       IND 226   15.0   39.1
18  WCLP CHATSWORTH GA    ETV 333   63.7   21.2

UCA-8
28  WJSP COLUMBUS GA      ETV 201   80.0    4.5

QCAUHF
30  WETV ATLANTA GA       ETV 223   11.9   37.0
36  WATL ATLANTA GA       IND 223   11.9   40.6
46  WANX ATLANTA GA       IND 223   11.9   37.8
```

# HEADEND APPLICATION NOTE

# ANTENNAS

Bob Mauney
Cable Communications Division
July 1979

In the late fifties, Government research produced a breakthrough in antenna state-of-the-art. This breakthrough was the concept of "frequency-independent" antennas and was based on the theory that if a structure is made proportional to itself by scaling of its dimensions by some ratio $\tau$, it will have the same properties at a frequency f and at a frequency $\tau$f. Therefore, the patterns and impedance of the antenna are periodic functions of the logarithm of frequency with a period of log $\tau$. By the proper choice of $\tau$, the properties of the periodic type of antenna will vary only slightly over the frequency band f to $\tau$f.*

## Antennas

A schematic diagram of a log-periodic dipole array is shown in Fig. 1. The antenna consists of parallel, linear dipoles arranged side by side in a plane. Practical considerations of construction, however, dictate that the dipoles do not lie exactly in a plane. The length of the dipole elements and the spacing between elements form a geometric progression. (The common ratio $\tau$ and taper angle $\propto$ are shown in the schematic.) The dipole elements are energized from a balanced constant-impedance feeder, with adjacent elements connected to the feeder in an alternating manner to obtain 180 degrees phase shift between elements. The antenna is fed by a coaxial line running from back to front through one of the support members, and the center conductor of the coaxial line is connected to the other support members. This type of connection forms an infinite balun, since the external portion of the antenna structure past the resonant element carries negligible current. Radiation from the antenna is end-fire in the direction of the decreasing elements.*

For satisfactory operation the antenna must contain a dipole at least 0.5 wavelength long at the lowest operating frequency and a dipole element shorter than 0.38 wavelength at the highest operating frequency. Theoretically, there is no limit to the bandwidth of a log-periodic antenna. The lower frequency cutoff is determined by the size of the antenna, while the upper frequency cutoff is determined by how accurately the elements are scaled. Scientific-Atlanta has made log-periodics with bandwidths in excess of 23 to 1.*

The performance of a frequency-independent antenna designed to operate over channels 7 through 13 (174 to 216 Mc) is summarized in Fig. 2. Note there are not dropouts in the performance; gain varies only 1 dB over the band. The E-plane beamwidths vary from 50 degrees to 54 degrees, and H-plane beamwidths vary from 60 degrees to 72 degrees. The front-to-back ratio varies from 25 dB to 35 dB. A typical E-plane pattern of this antenna is shown in Fig. 3 and a photograph in Fig. 4. Similar antennas covering channels 2 through 3 and 4 through 6 are also available.*

A common problem encountered by many systems is that of co-channel. The problem arises when the same channel from two different sources is received at a common point (Fig. 5). The CATV antenna picks up both signals and processes the two independent signals as if they were from a single source. If the two signals arrive from originating points which are separated in

bearing by approximately 30° or greater than the type of receiving antenna (antenna array) can be of great benefit in minimizing the degree of co-channel interference. By utilizing frequency-independent antennas as elements, an antenna array can be configured which has low sidelobes and therefore exhibits maximum rejection to the received signal producing the co-channel interference.

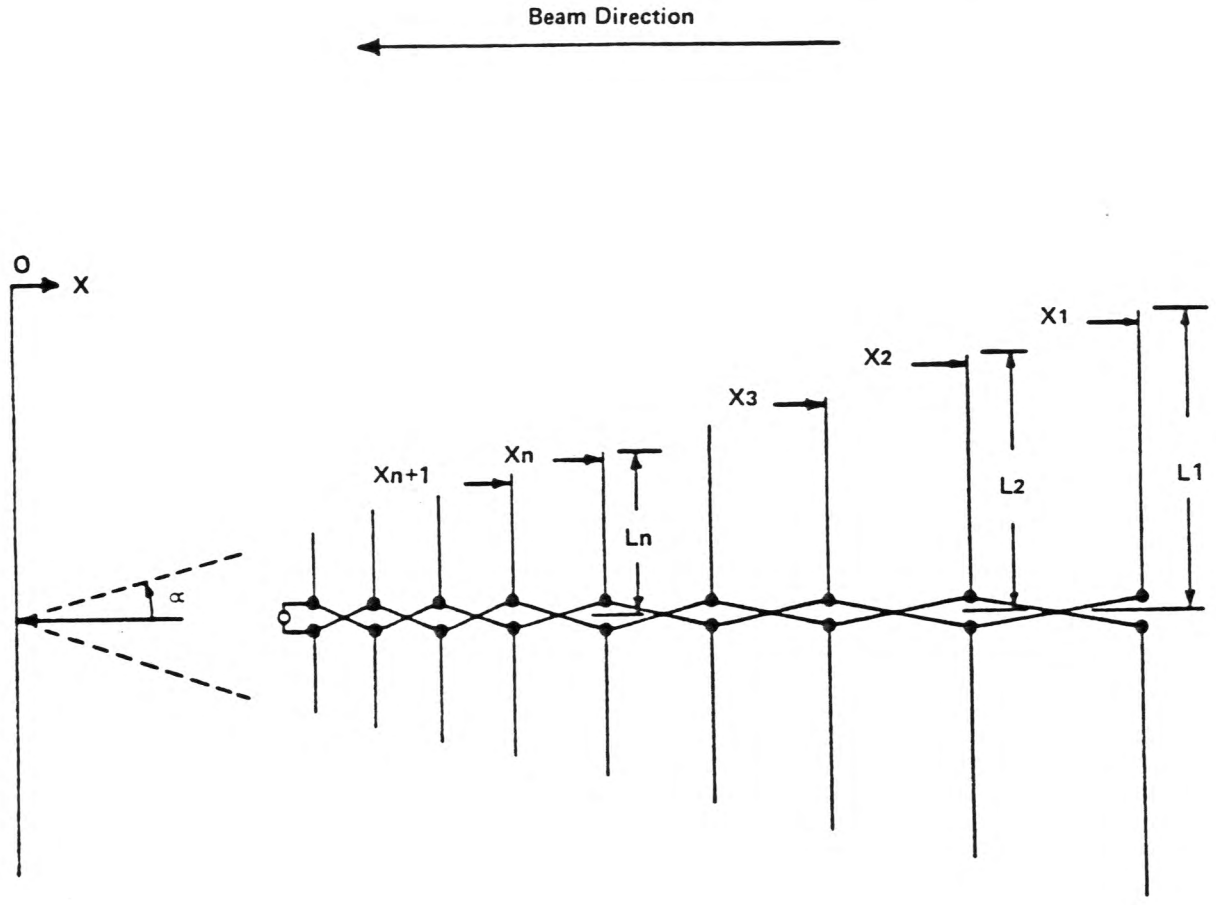



**Figure 1. Schematic Diagram of a Log-Periodic Array**

The type of antenna used to construct the array has a decided effect on the degree of rejection obtainable. If the E-plane pattern of a quad array of log periodic antennas is compared to that of a quad array of Yagi antennas, the advantage of the log periodic array is immediately obvious. As seen from the E-plane patterns, Fig. 6 and 7, the log array exhibits far superior rejection to

interfering signals arriving at bearing angles of 30° or greater. The difference between the rejection characteristics (E-plane patterns) of the log and the Yagi is even greater when the effect of the mounting structures are considered. Unlike the cantilever type mounting of the log and log array antennas, the Yagi and Yagi arrays must be supported in such a manner as to cause the supporting structure to be a decided influence upon the resulting pattern. The cantilever support of the elements of the log array eliminate the influence of the support tower and the E-plane pattern is not affected. A quad array of log-periodic antennas and the cantilever method of support is pictured in Fig. 8. A study on the effect of mounting structures of the Yagi and Yagi array antennas was undertaken and documented by Scientific-Atlanta's J.B. Weston, Jr., in June of 1966. The study is entitled, "Analysis of CATV Antenna Array Characteristics Utilizing Radiation Pattern Measurements" and is available upon request.

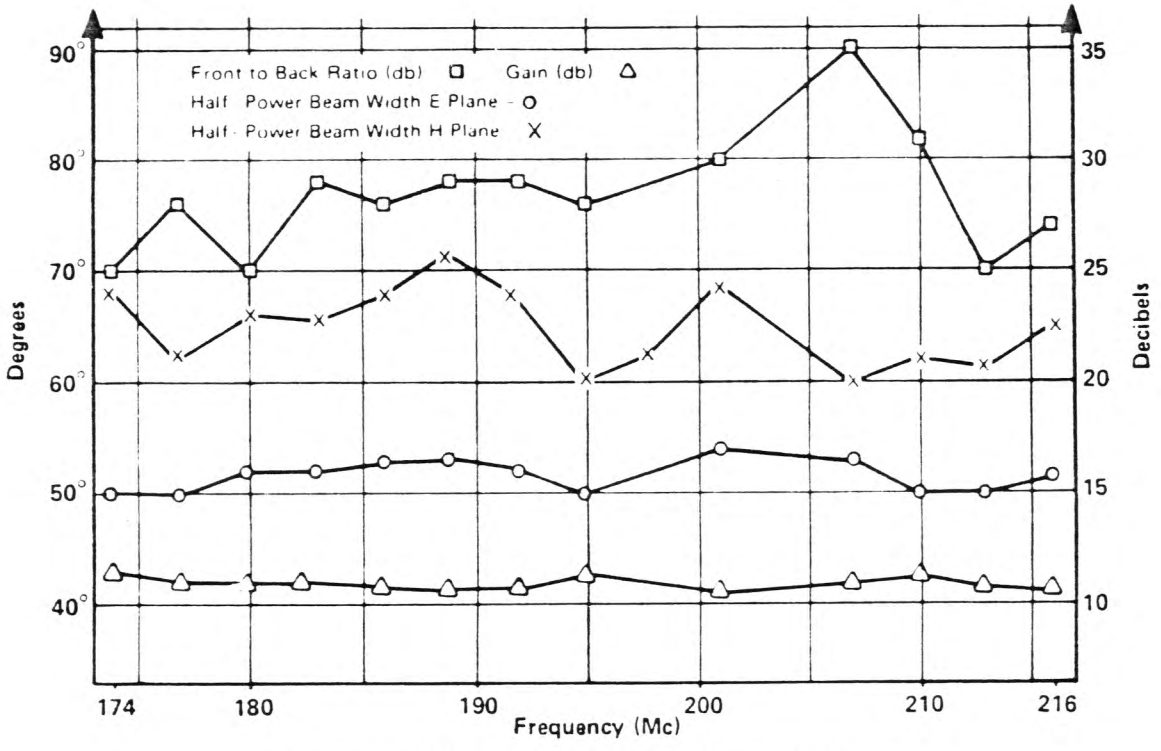



**Figure 2. Performance of Channel 7-13 Antenna**

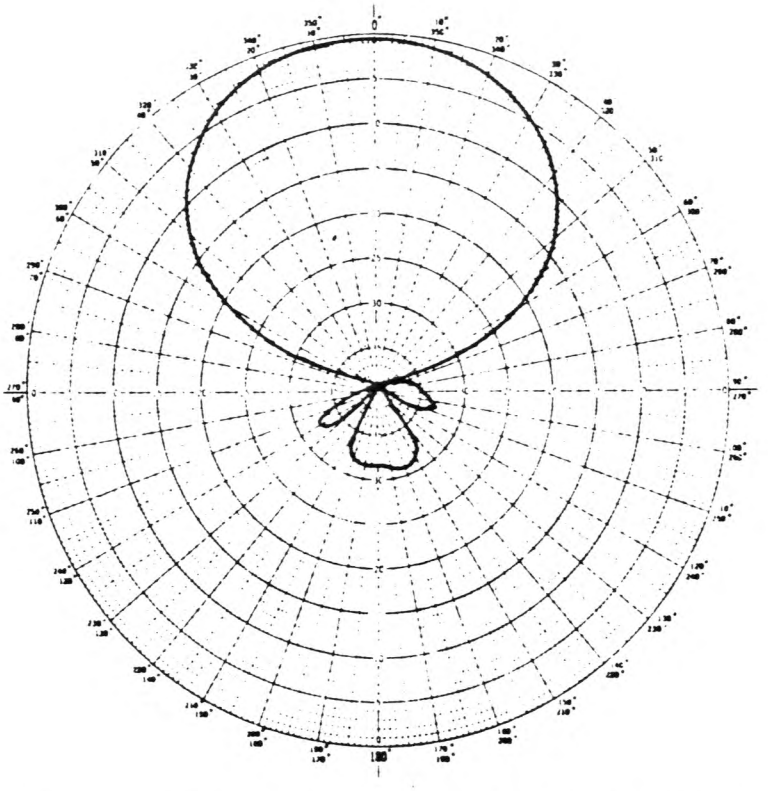


**Figure 3. Typical E-Plane Pattern of Channel 7-13 Antenna**

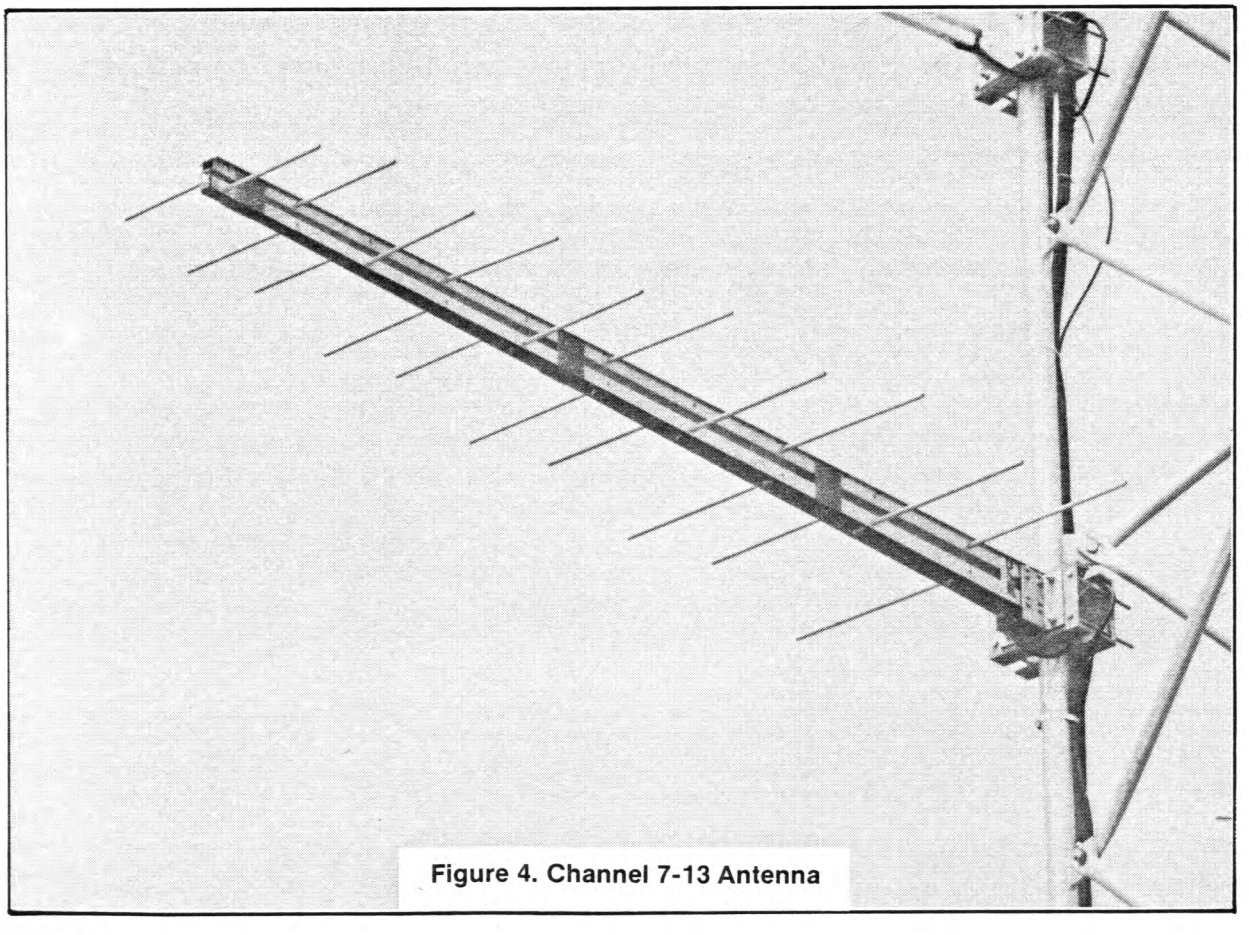


**Figure 4. Channel 7-13 Antenna**

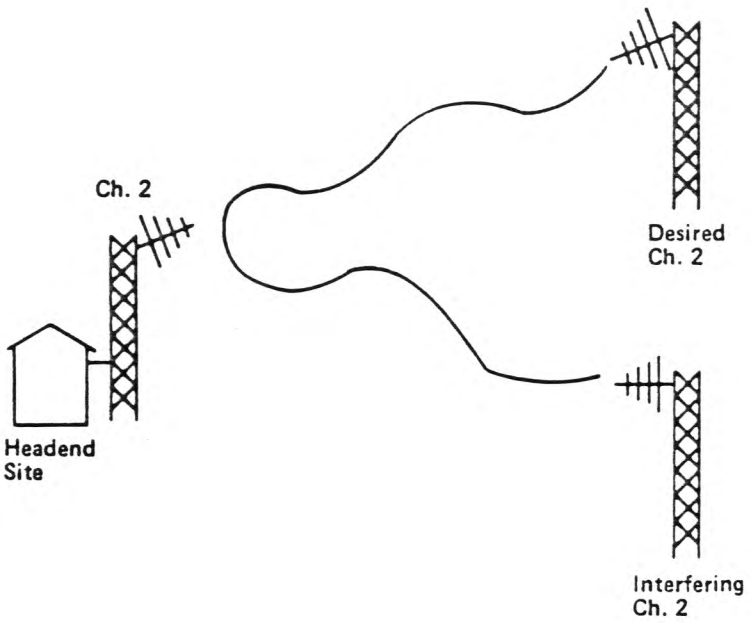



**Figure 5. Co-Channel Interference**

Degree of interference caused by the co-channel signal can be generally described as follows:

1. "Continuous" co-channel if difference in levels is $<$ 40 dB.
2. "Frequent" co-channel if difference in levels is 41 - 50 dB.
3. "Occasional" co-channel if difference in levels is 51 - 60 dB.
4. "No" co-channel if difference in levels is $>$ 60 dB.

An antenna derives its directive characteristics because of pre-determined and controlled spacing of the elements used in its construction. In order to avoid altering an antenna's characteristic pattern it is therefore necessary to consider the effect of close metallic objects including other antennas. This is usually of major concern in the initial tower layout for the headend. The following information can be used as a guide to determine the proper tower positions and spacing of Scientific-Atlanta antennas.

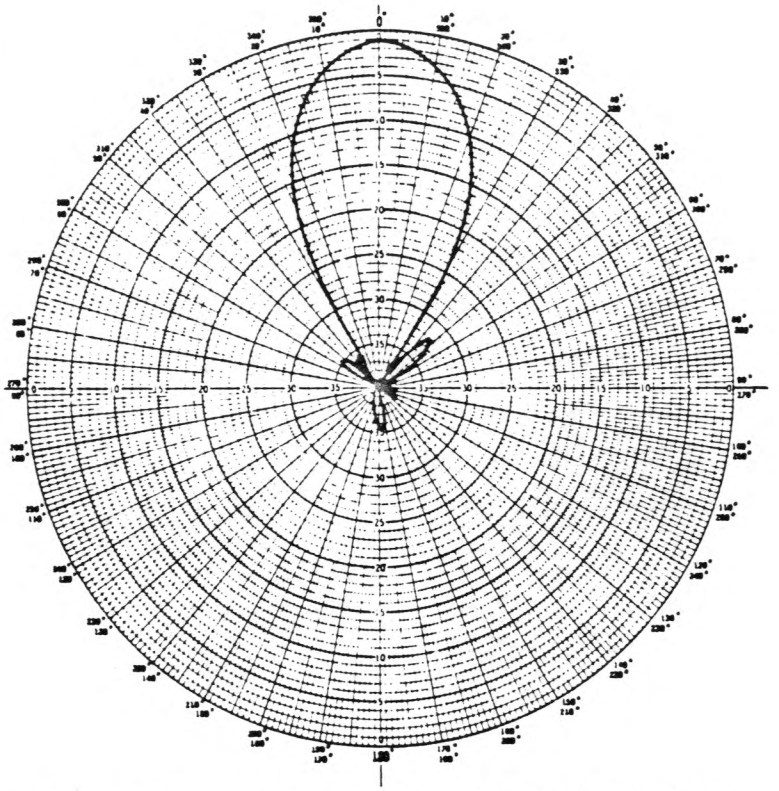


**Figure 6. Typical E-Plane Pattern of Quadrate Channeler**

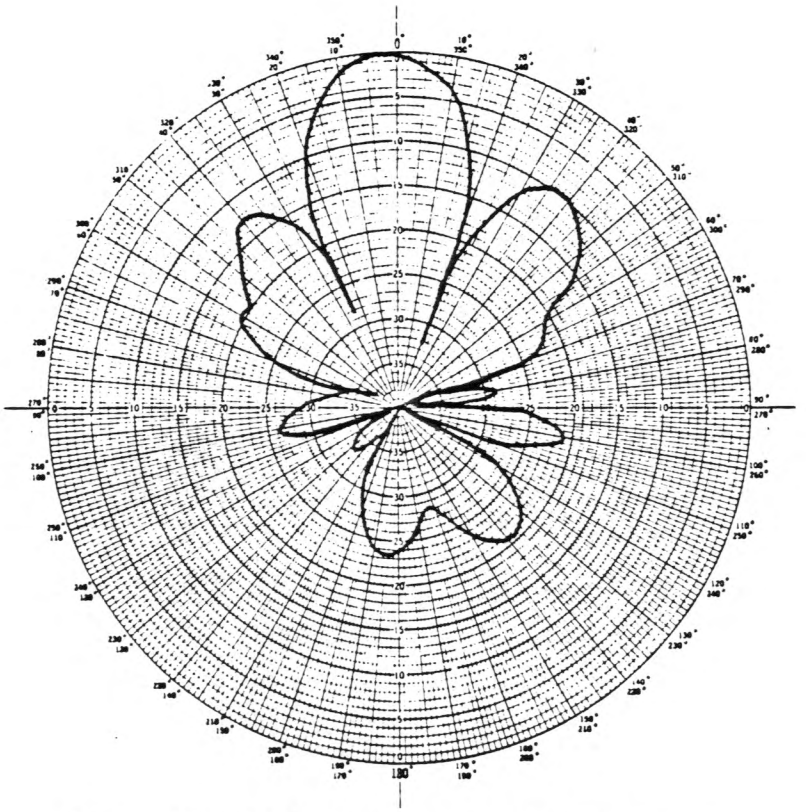
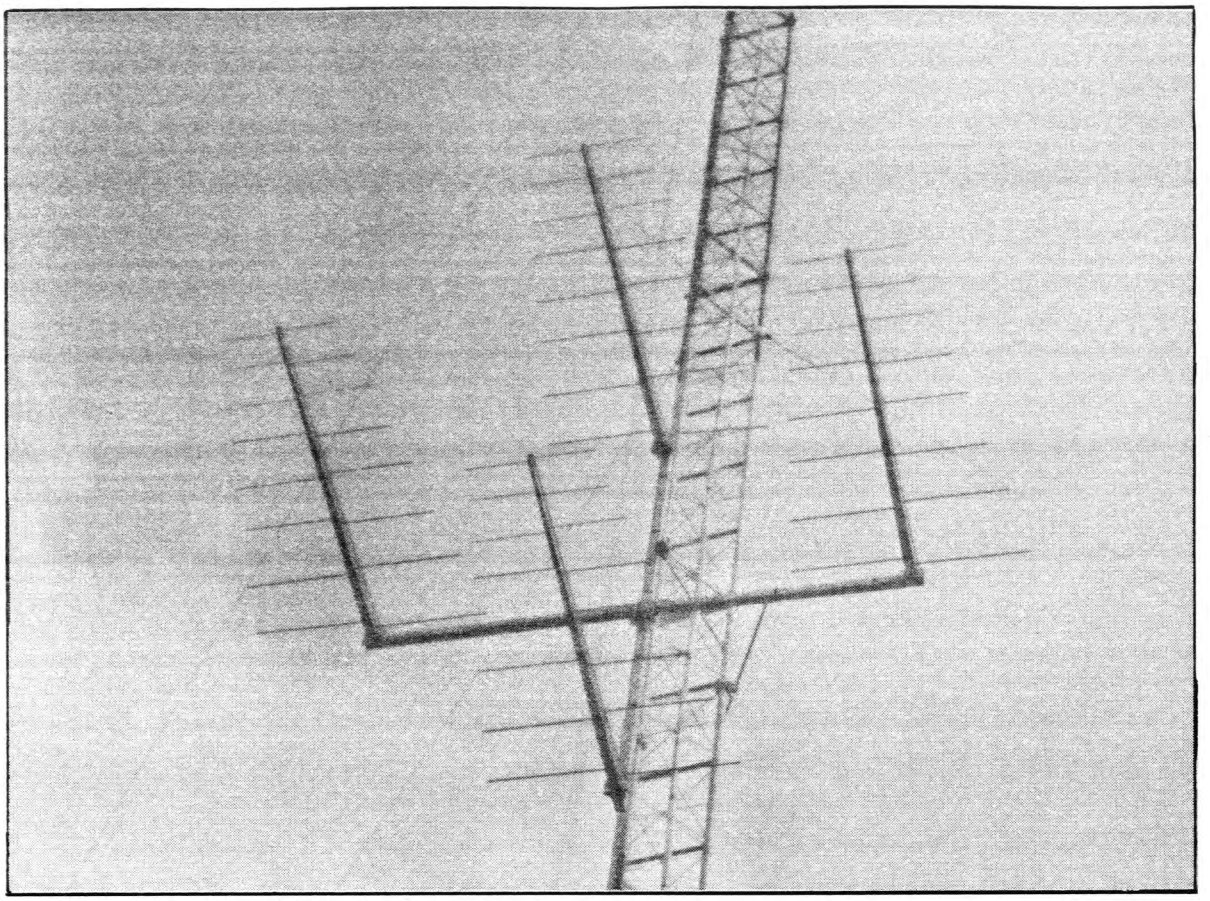


**Figure 7. Typical E-Plane Pattern of Quad-Yagi Array**



P2657

**Figure 8. Quadrate Channeler Antenna, Installed on Tower**

1. Antennas for reception of the weakest signals should be placed as high as possible on the tower.
2. If all received signal levels are approximately the same, then the highest frequency antennas should be placed toward the top of the tower.
3. For antennas located in a direct line below or above each other (same bearing) provide for a minimum spacing to the closest elements of one wavelength at the lowest frequency for which the antenna is designed (Table 1).

$$A = \lambda \text{ (feet)} = \frac{984}{\text{Freq. (MHz)}}$$

**TABLE 1**

| Antenna Model | A = Minimum Distance to Closest Metal Structure or Other Antenna Extremity | B = QCS Radi-Distance From Center Line to Outermost Extremity |
|---|---|---|
| QCA, QCS, QCD-2-6 | 18.22 Ft. | 8.75 Ft. |
| QCA, QCS, QCD-2 | 18.22 Ft. | 11.66 Ft. |
| QCA, QCS, QCD-4 | 14.91 Ft. | 8.75 Ft. |
| QCA, QCS, QCD-7 | 5.66 Ft. | 3.75 Ft. |
| QCA, QCS-UHF | 2.09 Ft. | 1.39 Ft. |
| QCA-FM | 11.18 Ft. | |

To calculate the required center line-to-center line mounting dimensions based on Item No. 3 refer to Table 1, Fig. 9 and the following relationships.

A. Center line-to-center line dimension for two QCS antenna arrays = A greater + $B_1$ + $B_2$.

B. Center line-to-center line dimension for one QCS and one QCA or one QCD antenna = A greater + B.

C. Center line-to-center line dimension for one QCA and one QCA or one QCD antenna = A greater.

Where

A greater is the greater of the two A dimensions as listed in Table 1 and B. $B_1$, $B_2$ are the individual radi of the arrays as listed in Table 1.

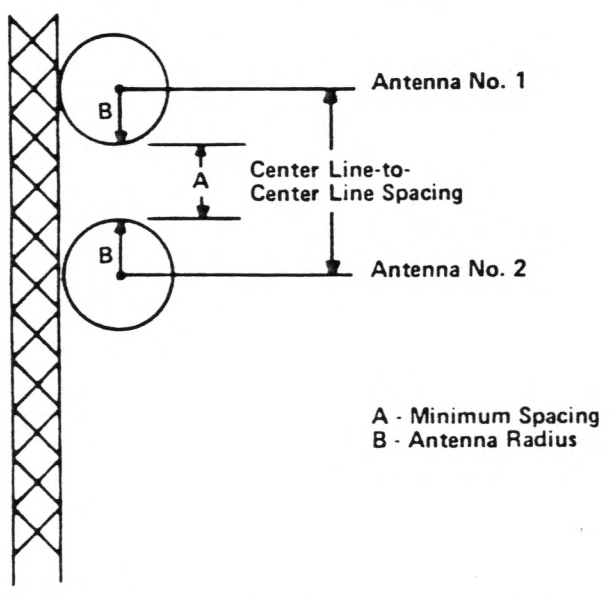



**Figure 9. Antenna Spacing for Same Bearing**

The above relationships and Figure 9 provide minimum spacing requirements for antennas positioned directly above or directly below each other. The same relationships can also be used

in cases where the bearing to the stations differ if the A dimension is modified as shown in Fig. 10 and calculated according to the following expression. The single qualification for calculating the spacing in the above manner occurs when $A_n$ goes to zero (i.e., when two antennas are displaced in bearing by 180°). Under this condition the two antennas may be mounted back-to-back with the minimum distance between center lines equal to zero.

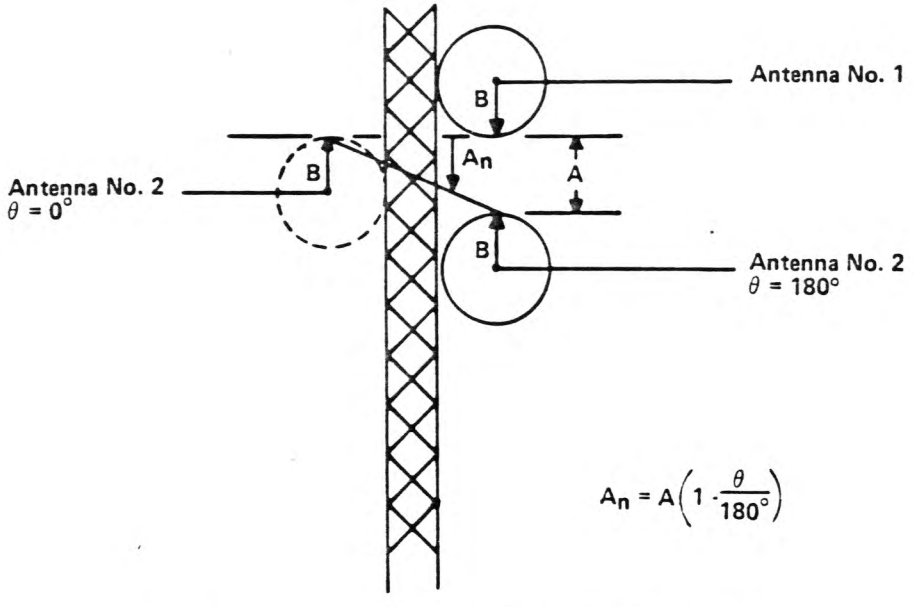



**Figure 10. Antenna Spacing for Bearing Displacement**

Where

A = Minimum spacing for antennas positioned directly above or below each other. (A = A greater when antenna models are not the same.)

$\theta$ = Angular Displacement of two antennas 0° - 180° (The difference in bearing between stations).

$A_n$ = New minimum distance to closest antenna extremity as a result of angular displacement between two antennas.

Scientific-Atlanta CATV antennas consist of two component main parts, (1) the antenna structure and (2) the feed line assembly. The antenna structure provides for specific electrical properties dependent upon fixed mechanical relationships between the component parts from which it is constructed. As a result of the mechanical nature of the antenna structure the electrical performance is fixed by final design and thus once manufactured can only be altered by changes to the basic mechanical structure itself. This of course assumes that the antenna is used properly and is spaced the appropriate wave length distance from other metal structures. If changes occur to the basic antenna structure such changes are usually the result of damage to

either the boom or more typically the antenna elements. If the antenna elements are broken or otherwise damaged repairs can be accomplished according to the following procedures:

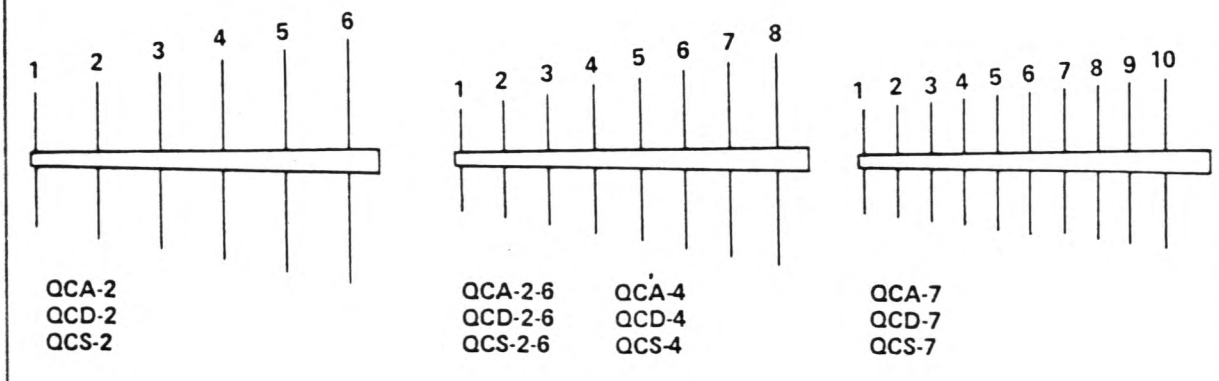
1. Obtain the proper replacement elements by referring to Table 11 and Figure 11.

TABLE II – Replacement Elements

| Element No. | Replacement Element Part No. | | | |
|---|---|---|---|---|
| | QCA/QCD/QCS-2 | QCA/QCD/QCS-4 | QCA/QCD/QCS-2-6 | QCA/QCD/QCS-7 |
| 1 | 55809 | 55813 | 55836 | 55820 |
| 2 | 55810 | 55814 | 55837 | 55821 |
| 3 | 55811 | 55815 | 55838 | 55822 |
| 4 | 55812 | 55816 | 55839 | 55823 |
| 5 | 55853 | 55817 | 55844 | 55824 |
| 6 | 55855 | 55818 | 55845 | 55825 |
| 7 | | 55819 | 55857 | 55826 |
| 8 | | 55851 | 55859 | 55827 |
| 9 | | | | 55828 |
| 10 | | | | 55829 |

**Installation instructions are included with each replacement element.**

**When ordering replacement elements for your antenna or array, include the following information:**
**(a) Antenna or array model number (from nameplate)**
**(b) Element number (counting from front-see below)**
**(c) Replacement element part number (from table below)**




QCA-2
QCD-2
QCS-2

QCA-2-6
QCD-2-6
QCS-2-6

QCA-4
QCD-4
QCS-4

QCA-7
QCD-7
QCS-7

**Figure 11. Antenna Element Numbers**

2. Remove the broken element. Saw or file any remaining burrs flush with the support channel. Do not remove the section of the old element that is inside the support channel.
3. Insert the replacement element through the section of old element inside the support channel.

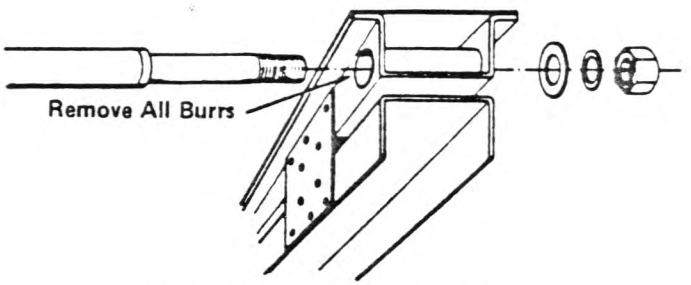



**Figure 12. Support Channel**

4. Install the flat washer, lockwasher, and nut as shown above. The flat washer must be next to the support channel.
5. Tighten the nut to secure the replacement element. Do not overtighten and deform the support channel.

Once the replacement elements are in place and secured the full electrical performance of the antenna should be restored.

The purpose of the mechanical components of the antenna structure is to gather and concentrate signal energy. Once this function is achieved, it is then necessary to provide a transportation medium to transfer the received concentration of energy from the antenna structure to a device for processing. The component of the antenna responsible for this task is the feed line.

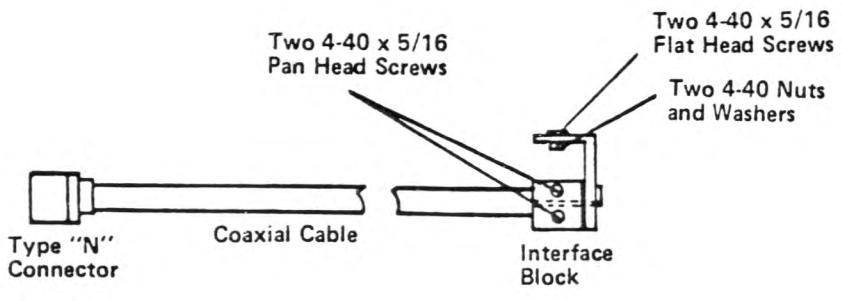



Figure 13. Antenna Feed Line

The feed line, like the antenna structure, is a mechanical assembly made up of coaxial cable, a connector and an interface block. The connector provides an easily accessible output from the antenna, the coaxial cable provides the transportation medium, and the interface block provides a means to couple the coaxial line to the antenna structure. This assembly is pictured in Fig. 13.

The feed line provides the interface to the antenna structure and therefore its electrical performance directly affects the overall performance of the antenna system. Because of the flexible construction of the coaxial cable the feed line is more susceptible to time related exposure to the environmental elements and therefore should be carefully inspected if a reduction in antenna performance is suspected. If it becomes necessary to replace the feed line the following procedure should be followed:

1. Obtain the proper feed line replacement from the factory.
2. Remove the feed line interface block from the narrow end of the antenna boom. Save the mounting hardware as it is removed.
    A. Remove four 4-40 x 5/16 pan head screws (two on each side) from the outside of the bottom boom channel.
    B. Remove two 4-40 x 5/16 flat head screws and nuts from the inside of the top boom channel. This step requires a screw driver for the 4-40 screws and an ignition wrench for the 4-40 nuts.
    C. Remove the interface block.
    NOTE: The area around the interface block has been coated with Scotch Cast electric rosin to prevent water from entering the end of the coaxial cable. It will be necessary to break this seal to remove the screws, nuts, and interface block. The electric rosin seal is not strong and does not require a great deal of pressure to break loose.
3. Cut the tie wraps holding the coaxial cable to the boom and remove the feed line assembly from the antenna.
4. Clean the contact area of the top boom channel with fine sand paper to ensure good mechanical and electrical contact between the channel and the new interface block.
5. Install the new interface block using the hardware removed in Step 2.
6. Tie wrap the coaxial cable to the bottom boom elements.
7. Coat the interface block area with electric rosin or some similar electrical insulating medium (such as RTV) to ensure that no water enters the end of the coaxial line.

Proper operation of the antenna should now be restored.

A note of caution is in order concerning replacement of feed lines on antenna arrays. The pattern characteristics of the array depend on all feed lines being the exact same length. Therefore when replacing a single feed line be certain that the length of the replacement perfectly matches the length of the feed line replaced.

*Taken directly from: "Basic Antenna Array Theory, Discussed with Respect to Minimizing Co-Channel Interference" by T.D. Smith (Scientific-Atlanta, Inc., 1965).

# SECTION 3
# 6150 CATV
# SIGNAL PROCESSOR

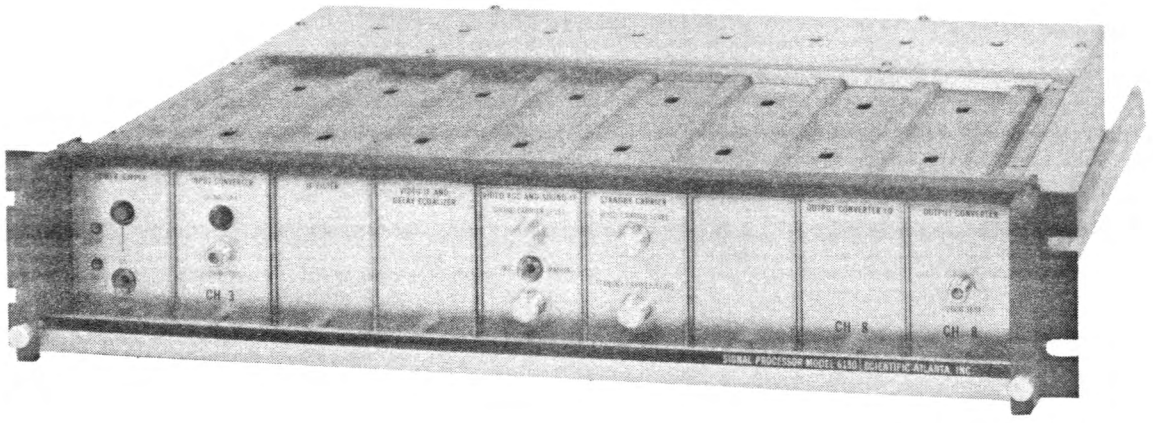


**Figure 3-1. Model 6150 Signal Processor**

## Description

The 6150 Signal Processor is totally solid state. Each circuit module is completely enclosed and is accessible from the front panel of the unit which occupies only 3-1/2 inches of vertical rack space. The modules of the processor consist of the power supply, input converter, IF filter, video IF and delay equalizer, video AGC and sound IF, standby carrier, output converter local oscillator, and the output converter. For the phase lock option, the output converter local oscillator and a blank module are replaced with the phase lock output converter local oscillator and the phase lock servo module.

Through the use of double balanced mixers and hybrids circuits spurious outputs are virtually eliminated from 5 MHz to 300 MHz.

Adjacent channel rejection is reduced to 60 dB minimum through the use of double-notches at each of the nearest adjacent channel carriers (upper adjacent channel picture carrier, lower adjacent sound carrier, and lower adjacent color sub-carrier).

Envelope delay is held to a minimum through the use of an all-pass phase equalizer located in the video IF amplifier.

The standby carrier oscillator has several modes of operation including CW, internally modulated with a 15 kHz keying pulse, and an external mode. This external mode greatly facilitates emergency audio and/or video override for community alert requirements. It also allows the operator to utilize his headend as a source of modulated signals with which to measure the cross-modulation in the distribution system.

Output frequencies from sublow through superband are standard. Input frequencies from sublow to UHF are standard.

## Options

- Low Level IF Switching (LIFS)
- High Level IF Switching (HIFS)
- Automatic IF Sensing (AIFS)
- Audio IF Override (AO)
- IF Loop Thru with Utility Video for Scrambling (IFLVO)
- Spectrum Inversion (SI)
- Modulation of Standby Carrier with Baseband Audio (IM)
- AGC Control on Alternate IF Input Provides for a Constant Output Level when the Alternate IF is Being Used (IFAGC)
- Phase Lock of the Output Picture Carrier is Available (PL)
- Standby Battery Powering Provides Backup for 115V ac (SP)
- Negative dc Powering Permits Operations from a −19 to −31V dc Battery Supply in Lieu of 115V ac (DC)

# Signal Processor Model 6150

## Specifications

Overall Sensitivity
100 Microvolt (−20 dBmV); Input for +60 dBmV output
Input Frequency Range
Any standard VHF, UHF, sub-low, mid-band, super-band IF, or any non-standard offset frequency
Input Level Dynamic Range
−20 dBmV to +30 dBmV with adjacent channels at the same level
Input Impedance and VSWR
75 — VSWR less than 1.35:1 over entire dynamic range for 6 MHz channel of interest
Output Impedance and VSWR
75 — VSWR less than 1.25:1 over 6 MHz channel of interest
Noise Figure
6 dB at max gain — Low Band
7 dB at max gain — High Band
9 dB at max gain — UHF
Frequency Response
±.5 dB from .75 MHz below to 4.2 MHz above video carrier
Envelope Delay from Picture Carrier to Color Sub-Carrier
Flat, ±25 nanoseconds
AGC Regulation
±.5 dB maximum output variation with input level variation from −20 dBmV to +30 dBmV
Output Frequency
Any standard VHF channel; sub-low; mid-band, super-band. Transposed video-sound channels available with optional spectrum inverter. Incremental and harmonically related coherent outputs are available, as are non-standard offset frequencies
Adjacent Channel Rejection
60 dB minimum
Sound Limiting
10 dB (min) at −25 dBmV input
Standby Carrier
Delay On — 10 sec ±3 sec
Delay Off — less than 1 sec
(Delay On can be internally programmed for instant On operation)
Frequency Accuracy of Crystal Oscillators
±.0025% from +20°F to +120°F for VHF
±.001% for UHF
Output Level
+40 dBmV to +60 dBmV continuously variable
Spurious Outputs
60 dB below video carrier with video carrier set at +60 dBmV and sound set at +45 dBmV
Output Level Range for Video and Substitution Carriers
Continuously variable from +45 dBmV to +60 dBmV
Output Level Range for Sound Carrier
Continuously variable from 10 dB to 20 dB below picture carrier

**GENERAL**

Input Voltage
100 to 130V ac, 50 Hz to 60 Hz
+24V dc nominal standby battery with optional built-in automatic switchover
−24V dc, nominal battery operation with optional dc converter module
Input Power
30 watts maximum
Weight
10.4 kg (23 pounds)
Mounting
89mm (3½")H x 483mm (19")W x 495mm (19½")D (Standard rack mount spacing; rack slides optional)
Operating Temperature Range
−7 to +50°C (+20 to +120°F)

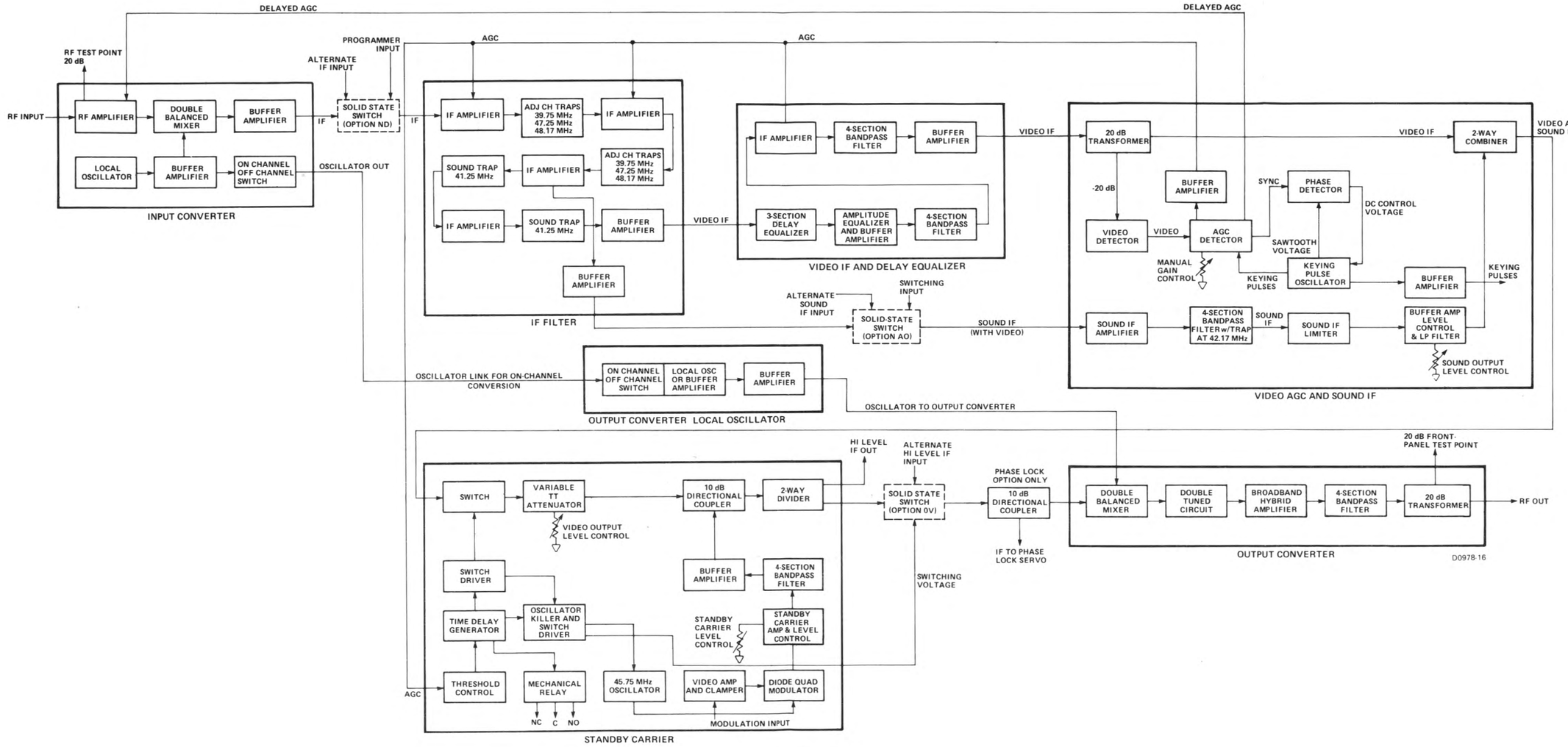



**Figure 3-2. Model 6150 and 6150PL Signal Processors, Detailed Block Diagram**

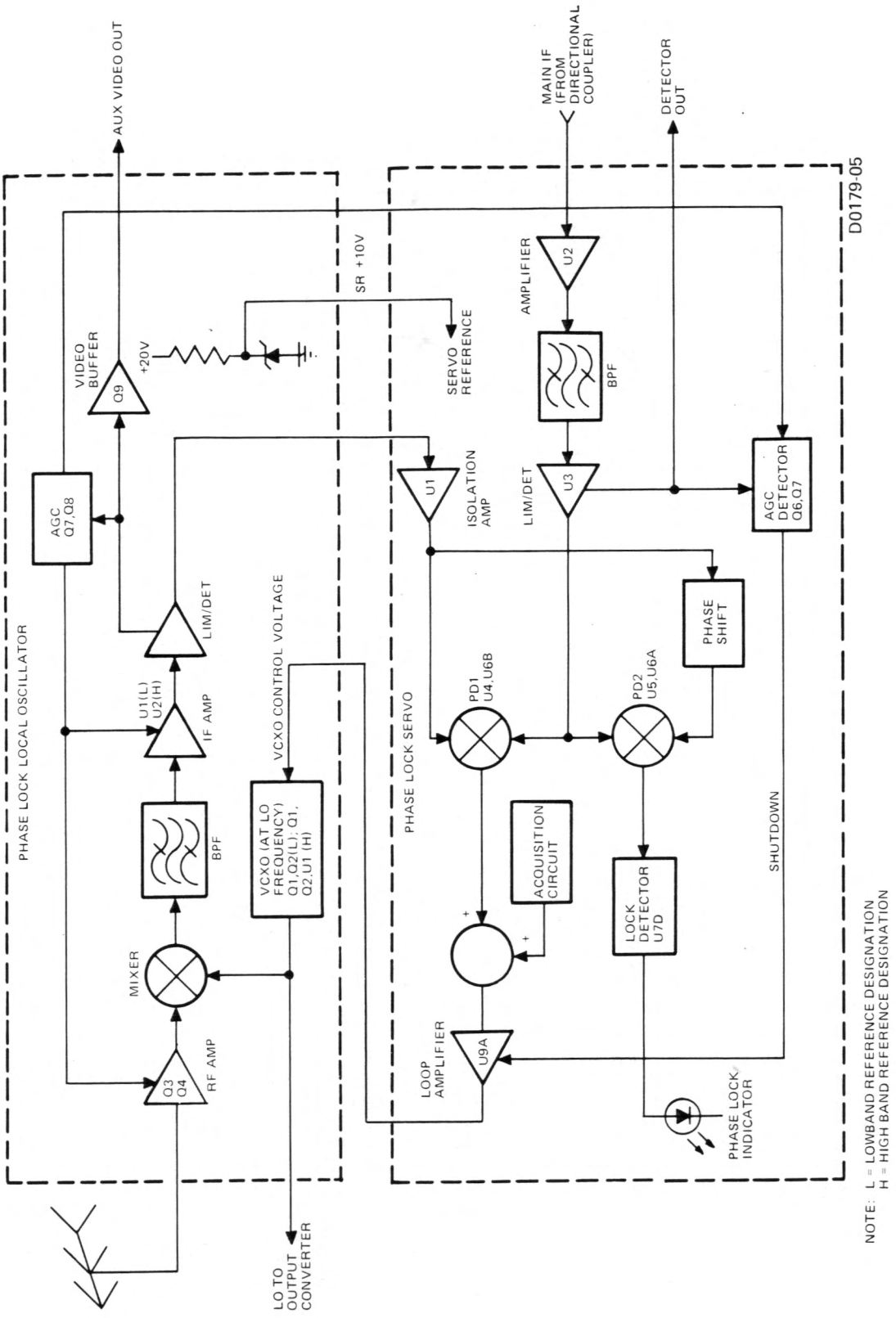
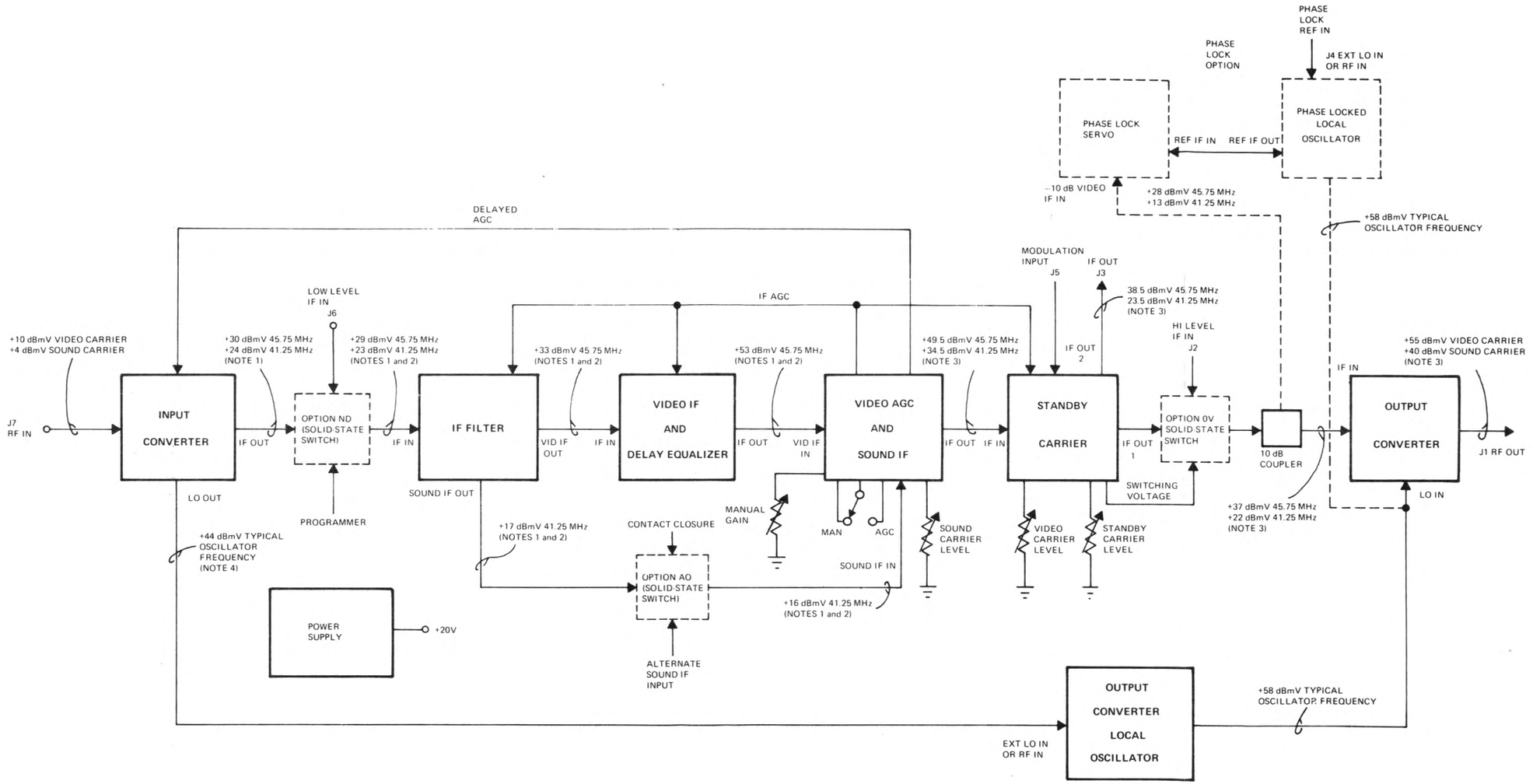



**Figure 3-3. Phase Lock Block Diagram for Low-Band Through Super-Band**




NOTES:

1. Before removing jumper cable, place gain mode switch in manual position and adjust for an output video carrier level of +55 dBmV. If no output is present, measure AGC voltage on rear apron test point No. 1 in AGC mode with input signal present, then place gain mode switch to manual and adjust manual gain control for same AGC voltage on test point No. 1.
2. This measurement assumes solid-state switch is not activated.
3. For All measurements made after the video AGC module, the gain mode switch should be left in the AGC position.
4. The levels indicated here should only be present for on-channel conversions.

**Figure 3-4. Model 6150 and 6150PL Signal Processors, Block Diagram**

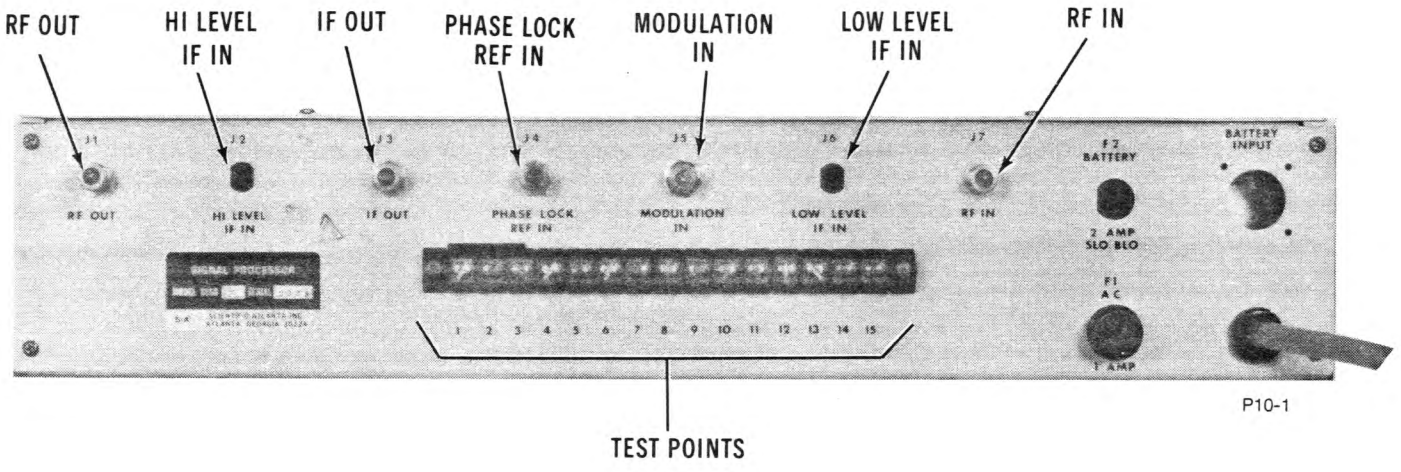



**Figure 3-5. Model 6150 Rear View**

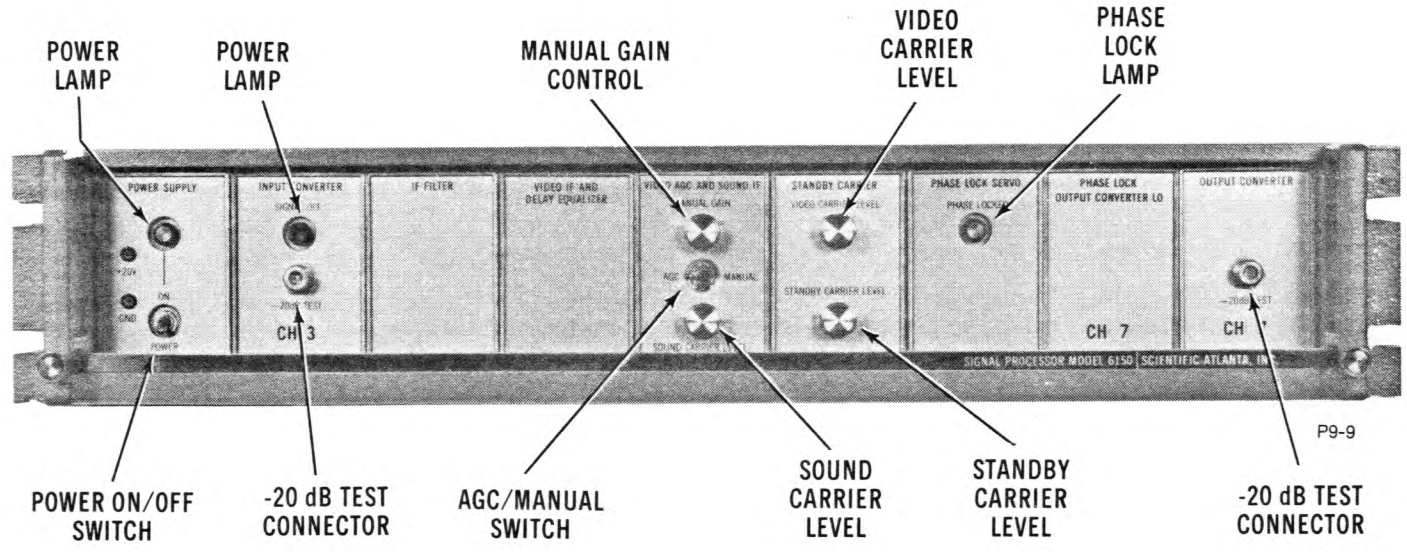



**Figure 3-6. Model 6150 Front View**

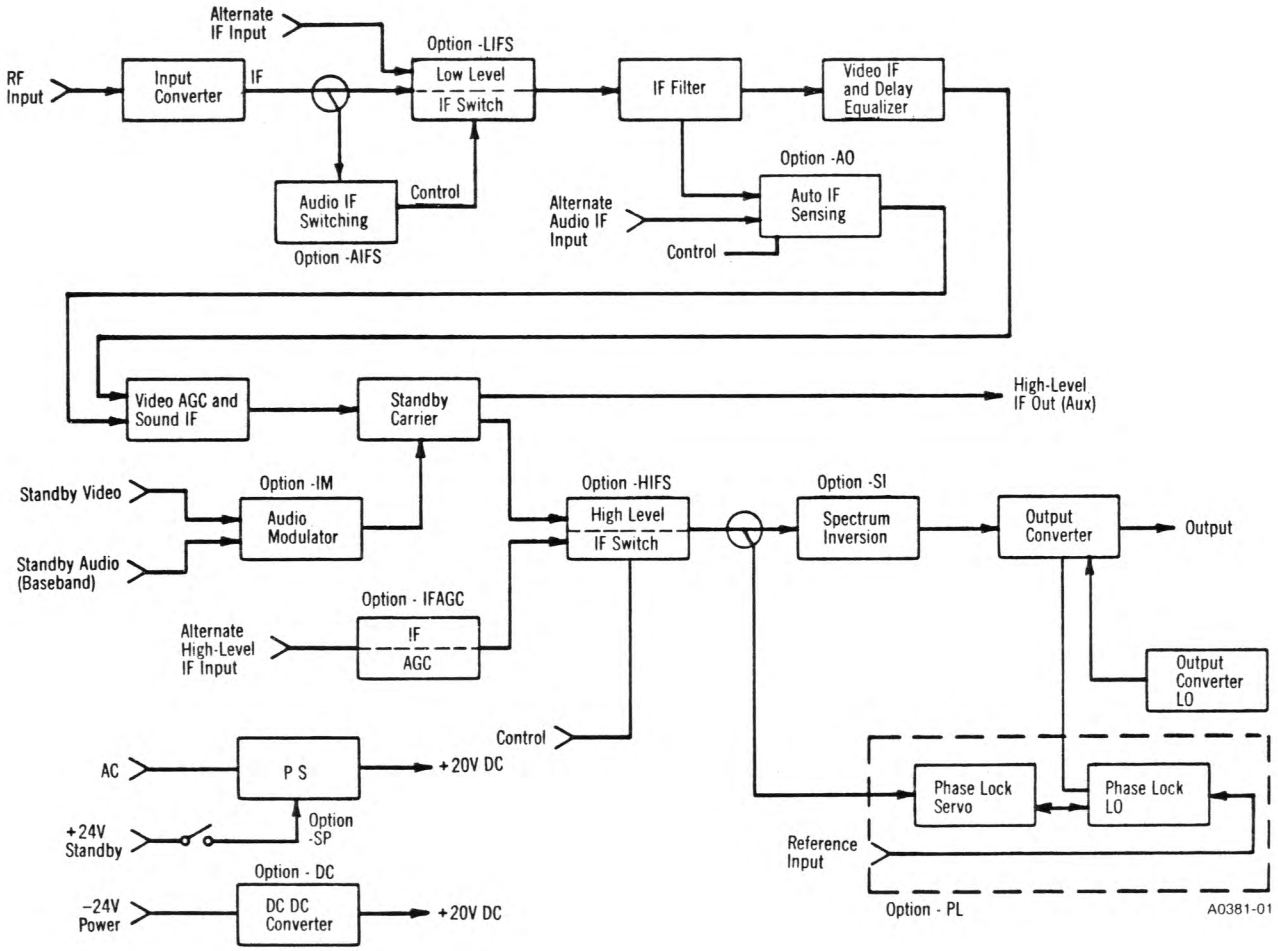



**Figure 3-7. Model 6150 Block Diagram**

## SIGNAL PROCESSING REQUIREMENTS FOR MODERN CABLE SYSTEMS

**By Alex Best**
**Scientific-Atlanta, Inc.**
**Atlanta, Georgia**

The movement of cable TV over the past few years from the rural areas into the large metropolitan areas has brought with it new demands on the techniques of processing signals. To a large degree these changes in processing requirements have been dictated by the fact that in metropolitan areas, the best geographical location for receiving signals off-air is separated by some distance from the most convenient office and local origination site, and in most instances both of these are located some distance from the optimum point to distribute the signals. Headend electronics capable of operating in this environment without producing excessive degradation have increased the technical performance specifications of headend equipment, especially heterodyne signal processors. To investigate these increased performance demands, consider a "typical" headend system for a metropolitan area such as the one shown in Figure 1.

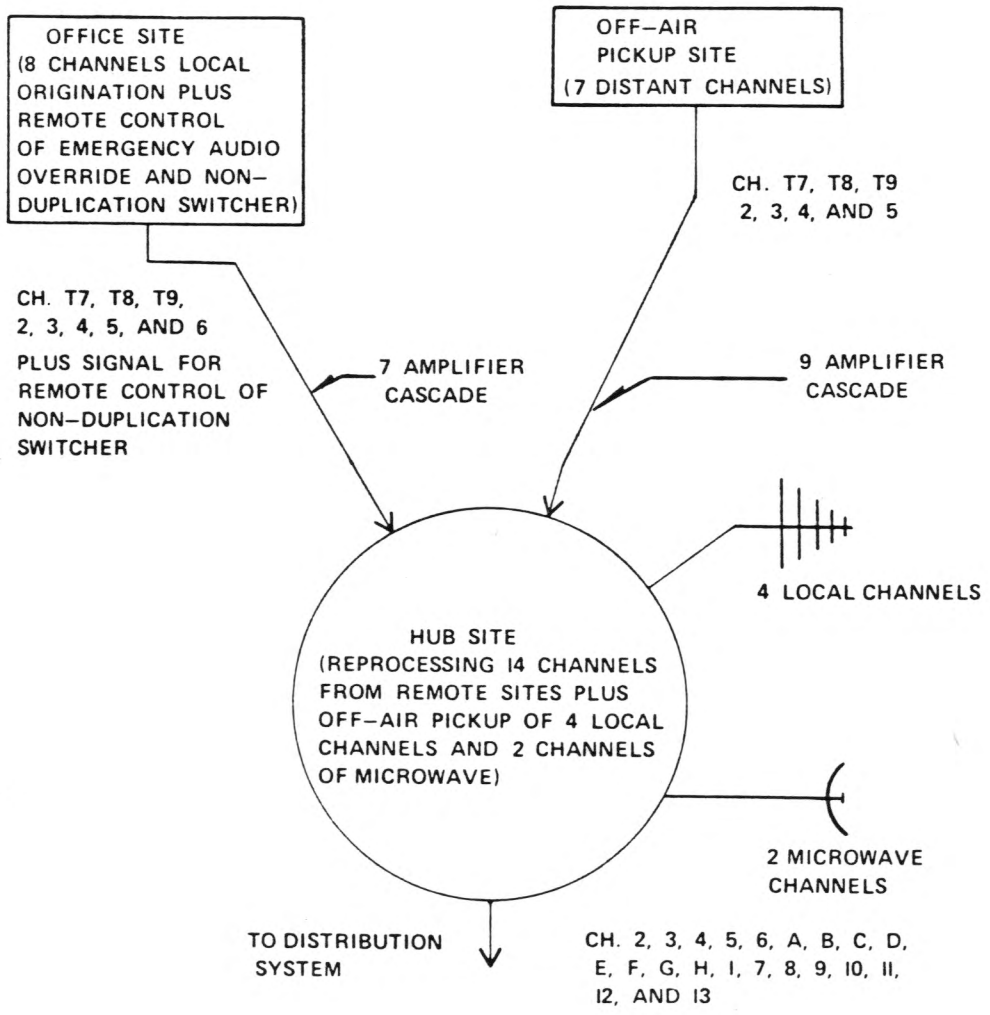

OFFICE SITE
(8 CHANNELS LOCAL ORIGINATION PLUS REMOTE CONTROL OF EMERGENCY AUDIO OVERRIDE AND NON–DUPLICATION SWITCHER)
OFF–AIR PICKUP SITE
(7 DISTANT CHANNELS)
CH. T7, T8, T9
2, 3, 4, AND 5
CH. T7, T8, T9,
2, 3, 4, 5, AND 6
PLUS SIGNAL FOR REMOTE CONTROL OF NON–DUPLICATION SWITCHER
7 AMPLIFIER CASCADE
9 AMPLIFIER CASCADE
4 LOCAL CHANNELS
HUB SITE
(REPROCESSING 14 CHANNELS FROM REMOTE SITES PLUS OFF–AIR PICKUP OF 4 LOCAL CHANNELS AND 2 CHANNELS OF MICROWAVE)
2 MICROWAVE CHANNELS
TO DISTRIBUTION SYSTEM
CH. 2, 3, 4, 5, 6, A, B, C, D, E, F, G, H, I, 7, 8, 9, 10, 11, 12, AND 13

**Figure 1. Typical Headend System for a Metropolitan Area**

Here we have the situation discussed above, where the office and local origination site, off-air pickup site, and hub site are physically located several miles apart. The processing problem arises at the hub site, particularly with the channels from the other sites that must be reprocessed. For the sake of discussion, let us consider the channel 3 signal being fed into the hub site from the off-air pickup site. This channel was originally received off-air as a standard VHF or UHF signal and converted to channel 3 in a signal processor. After combining with the sub-low and low band channels, it is fed through a 9 amplifier dedicated distribution system to the hub site. Here the combined signal is split as many as there are channels coming in and fed into the input of a second processor. In this discussion we will assume that the channel 3 coming from the off-air pickup site will be carried on channel 3 on the distribution system. This being the case the signals present at the input to the channel 3—3 processor at the hub site will appear as in Figure 2.

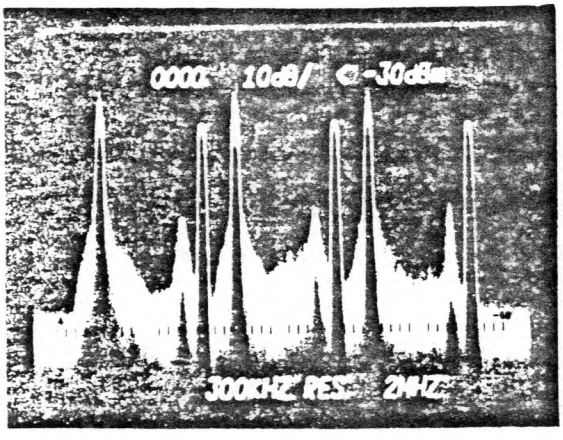


Figure 2. Ch 2, Ch 3, Ch 4 out of seven channels coming from off-air receiving site.

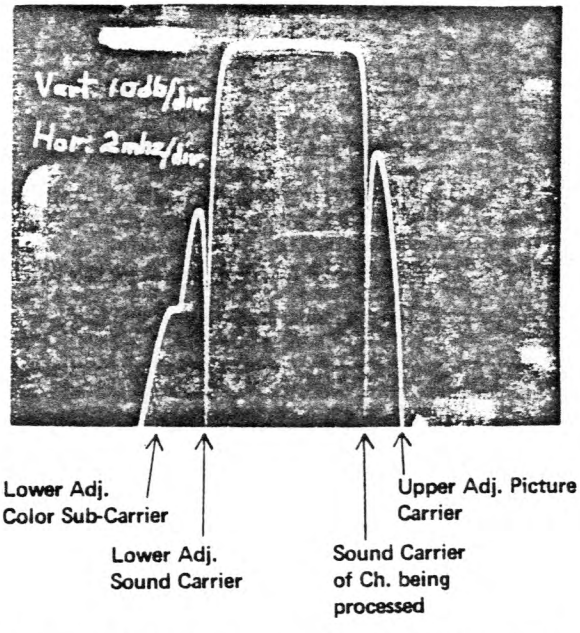



Figure 3. Amplitude Response of 6150 Signal Processor.

Here we show only three out of seven signals that are actually present. For purposes of demonstration the sound carrier levels have been set 6 dB below the video carrier levels when in actual practice they are set approximately 15 dB below. The question here is: how much adjacent channel rejection must this processor have in order not to interfere with other channels being distributed on the cable? The worst case problem would exist if the channel 4 signal being distributed on the system is different from the channel 4 signal present at the input to the channel 3—3 processor. Under these circumstances any of the channel 4 signal from the off-air pickup site feeding through the channel 3—3 processor would fall back on channel 4, with an unavoidable frequency difference between it and the channel 4 distributed on the system. In this case 60 dB would be the minimum adjacent channel rejection allowable. In Figure 3 we show the

amplitude response of a Scientific-Atlanta Model 6150 channel 3—3 processor. Here we see "notches" at the upper adjacent picture carrier frequency and lower adjacent sound carrier frequency in excess of 70 dB. We also see undesired but unavoidable "response upshoots" on the high frequency side about 20 dB down and on the low frequency side about 30 dB down. Would a processor with this amplitude response be specified as having 60 dB adjacent channel rejection? Shown in Figure 4 is the output spectrum of this processor with the input spectrum shown in Figure 2. Here we see that there is no energy falling in the upper and lower adjacent channels greater than 60 dB below the desired video carrier level. Therefore, if we define adjacent channel rejection as the rejection to equal level adjacent channels that have "normal" energy distributions (pc=0 dB, sc=-6 dB, cc=-17 dB), this amplitude response would have 60 dB adjacent channel rejection. The point here is that deep notches alone do not constitute adjacent channel rejection. Nor does "response upshoots" necessarily hurt you. It is a combination of the processor amplitude reponse and the energy distribution of a "typical" television signal that together make up adjacent channel rejection. Signal processors with 60 dB adjacent channel rejection have only recently become available in the CATV industry. Before this, processors located at hub sites required external traps/filters to achieve this degree of selectivity.

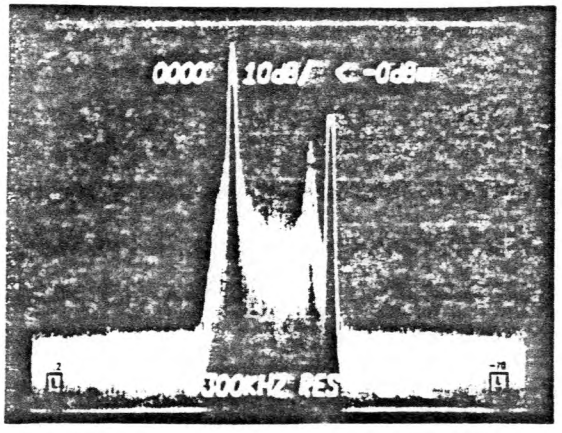



Figure 4. Output spectrum of signal processor with amplitude response shown in Figure 3 and input spectrum shown in Figure 2.

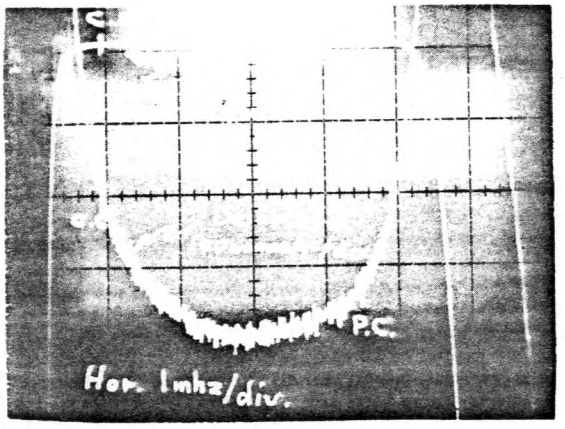



Figure 5. Amplitude and envelope delay response of IF section of 6150 processor (without delay equalization).

It should be pointed out that any additional increase in selectivity does not come without penalty. The price we pay is an increase in the differential time delay across the passband of the processor. Shown in Figure 5 is the amplitude and envelope delay response of the IF section of a Scientific-Atlanta 6150 processor, without delay equalization. The IF position of the processor constitutes the majority of the selectivity and therefore produces practically all the delay. In the CATV industry envelope delay is defined as the differential time delay between the picture carrier frequency and the color subcarrier frequency. From Figure 5 we can see this is approximately equal to 130 nanosec. for the processor with no equalization. It is also interesting to note that although the "envelope delay" is only 130 nanosec., the total differential time delay from band edge to the center frequency is approximately 500 nanosec. It should be pointed out that for "minimum phase shift" networks, into which category virtually all headend equipment falls, the amplitude response uniquely defines the delay response.[1] Therefore, regardless of how the response shown in Figure 3 is obtained, the delay would be the same.

To determine the detrimental effects of processing signals through a device that has the delay characteristics shown in Figure 5, the test set up shown in Figure 6 was used. Here we have a channel 3 modulator and channel 3 demodulator with the capability of either connecting them back-to-back or inserting the channel 3—channel 3 processor between them. Two video test signals were used in this evaluation, one being the "Modulated 20T pulse" and the other the "2T $Sin^2$ pulse." The first test signal was designed primarily to indicate the relative delay between the video carrier and the color subcarrier frequency. The "$Sin^2$ pulse" is more sensitive to the amplitude and delay in the vicinity of the response that the luminance signal energy occupies. Figure 7 shows the response of the modulator—demodulator back-to-back for the two test signals described above. As can be seen, the modulator—demodulator combination is very transparent, indicating very little distortion. The "Mcdulated 20T pulse" indicates virtually no envelope delay and the "$Sin^2$ 2T pulse" has a K factor of 2%.

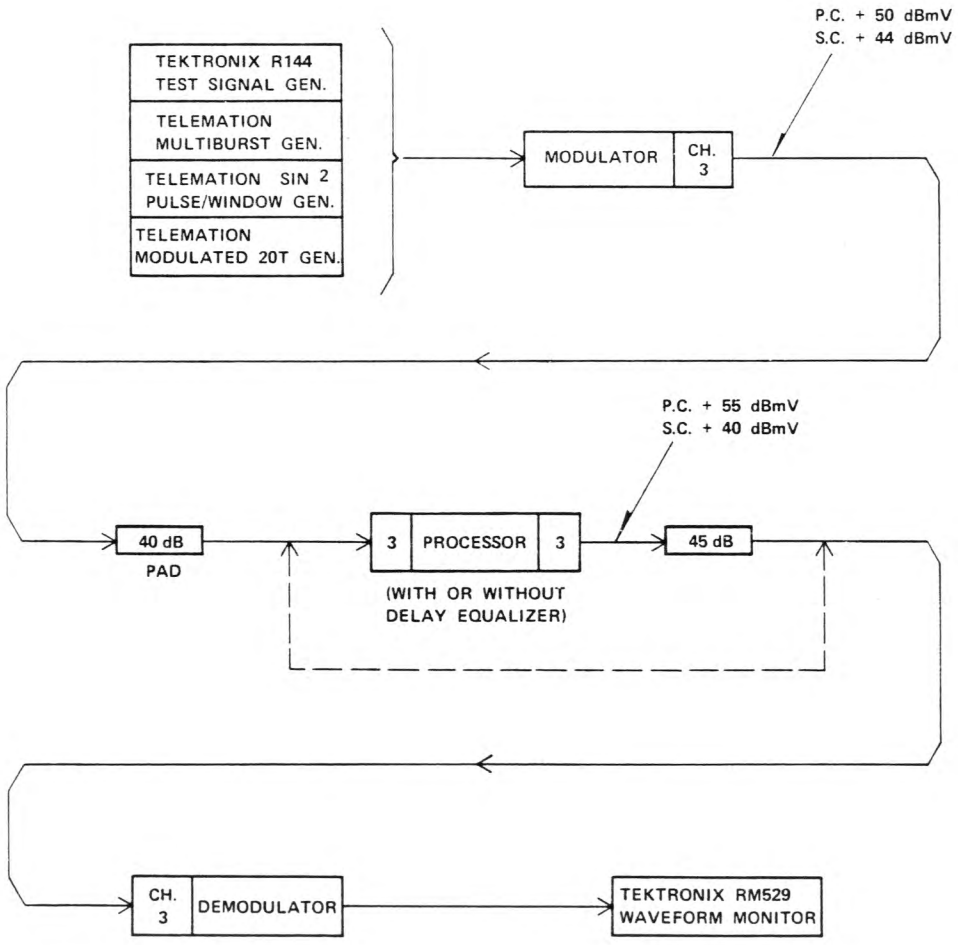



**Figure 6. Equipment Setup For Waveform Testing of Signal Processor.**

Shown in Figure 8 is the result of passing these test signals through the same mod-demod, but with a 6150 processor with no delay equalization inserted between them. The distortion to the 20T pulse indicates a delay to the color subcarrier relative to the video carrier of approximately 110 nanosec.[2] This measurement verifies the results found using the General Radio 1710 network analyzer. The distortion to the Sin$^2$ 2T pulse shows a leading undershoot and trailing smear which is characteristic of nonlinear envelope delay in the vicinity of the video carrier.[3] The K factor under these conditions has increased to 3.5%. This increase in distortion has taken place with the signal passing through only one processor with no delay equalization. In headend-hub systems many signals pass through two processors. In this case the envelope delay would become approximately 220 nanosec. and the K factor would increase by another 1.5%. This degree of distortion could lead to a trailing smear or "halo effect" on television images, especially those that have passed through two non delay-equalized processors. It should be pointed out that virtually all of the degradation we have seen is due to the nonflat envelope delay across the system passband. As we will see later, the amplitude response has very little amplitude variation. Also, there is no quadrature distortion since the demodulator used was of the synchronous detector type.

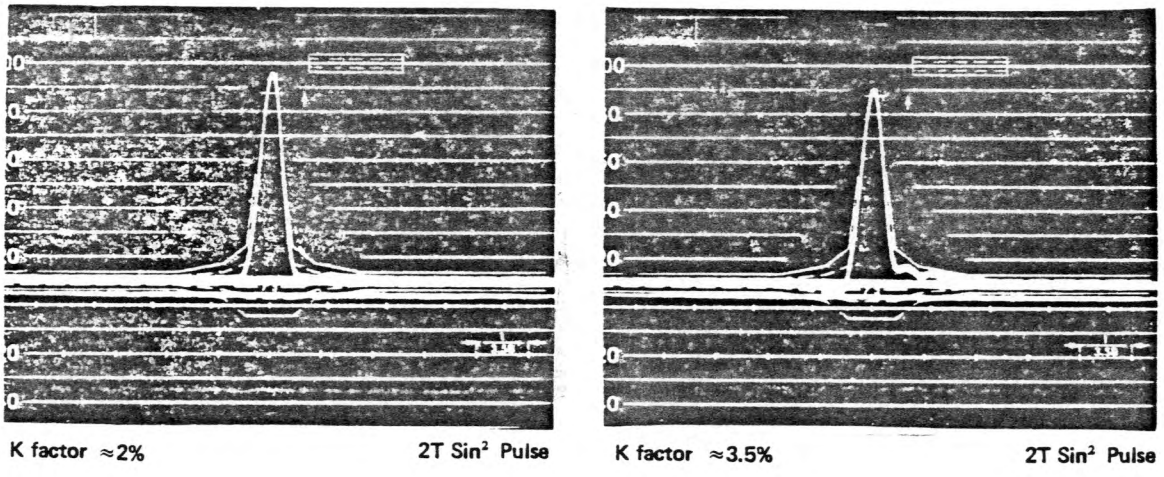



**Figure 7. Video Test Signal Reponse of Modulator — Demodulator.**

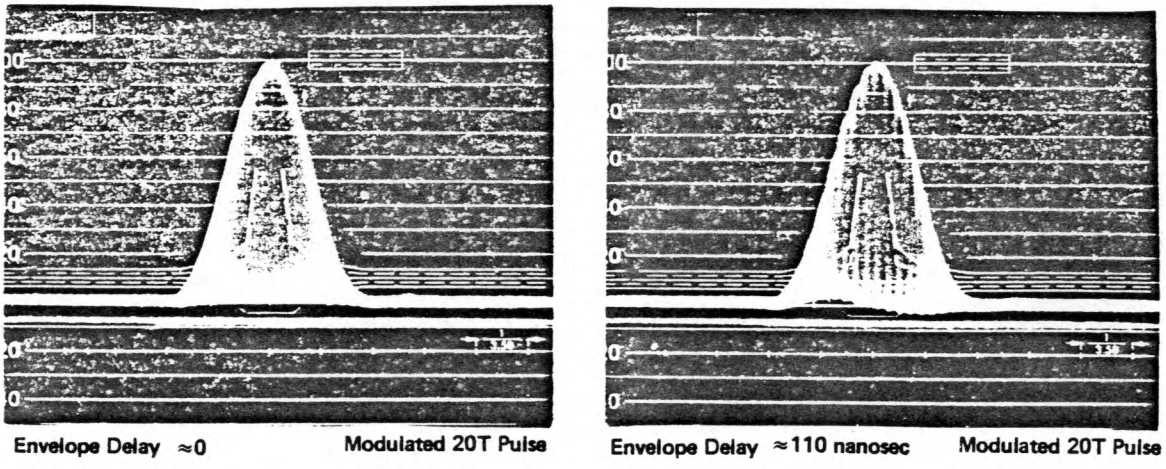



**Figure 8. Video Test Signal Response of Modulator — Processor (Without Delay Equalization) —Demodulator.**

Let us now examine the improvement that can be achieved by delay equalizing the signal processor. Figure 9 is a photograph of the amplitude and envelope delay response of the IF section of a Scientific-Atlanta 6150 processor, with delay equalization. From the amplitude response we see that although the all-pass delay equalizer is theoretically a lossless device, it does introduce a small amplitude variation (±.3 dB typical). Actually, the delay equalizer alone produces a 5 or 6 dB dip in the response which must be amplitude equalized to achieve the desired flatness. The results of passing the multiburst test signal first through the mod-demod back-to-back and then through the mod-processor (with equalizer)—demod can be seen in Figure 10. As evident from this photograph, the amplitude variation is certainly tolerable.

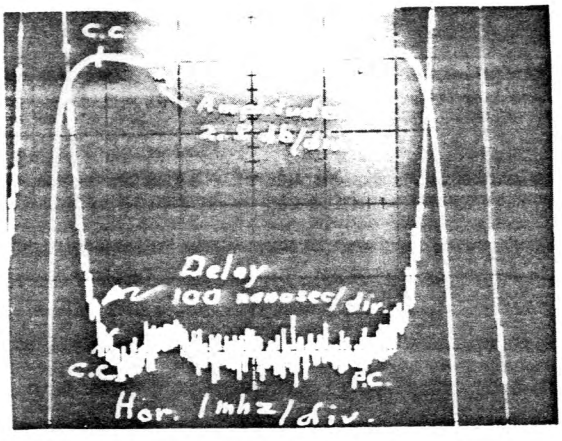
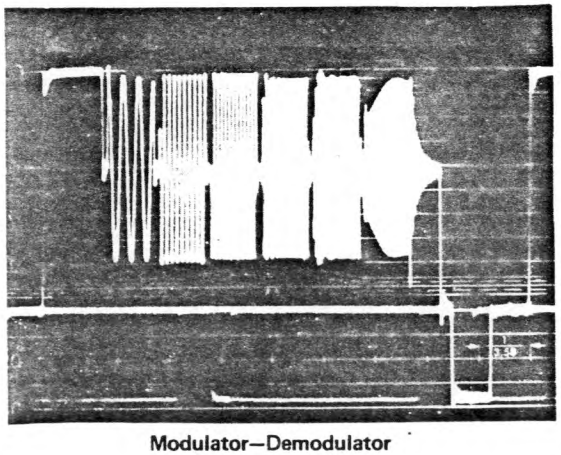
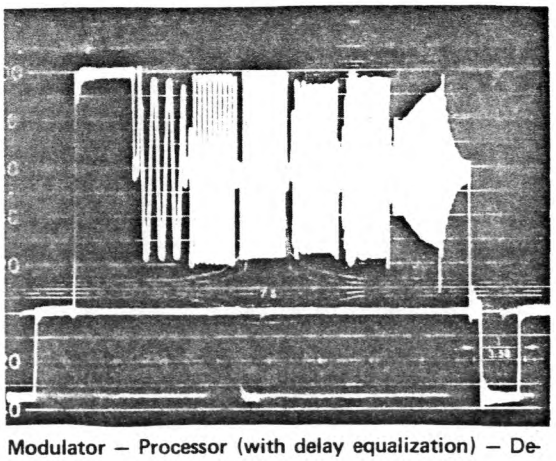



**Figure 9. Amplitude and envelope delay response of IF portion of 6150 processor (with delay equalization).**

Obviously, when comparing Figure 9 and Figure 5, the most dramatic difference caused by the delay equalizer is in the flatness of the envelope delay response. From this response it can be seen that the relative delay between the video carrier frequency plus .5 MHz and the color subcarrier frequency is flat ±20 nanosec. Also evident is that from the two frequencies mentioned above to the band edges the delay is approximately 200 nanosec. The reason for this compromise in equalized delay is two-fold: first, the increase in complexity in an equalizer to



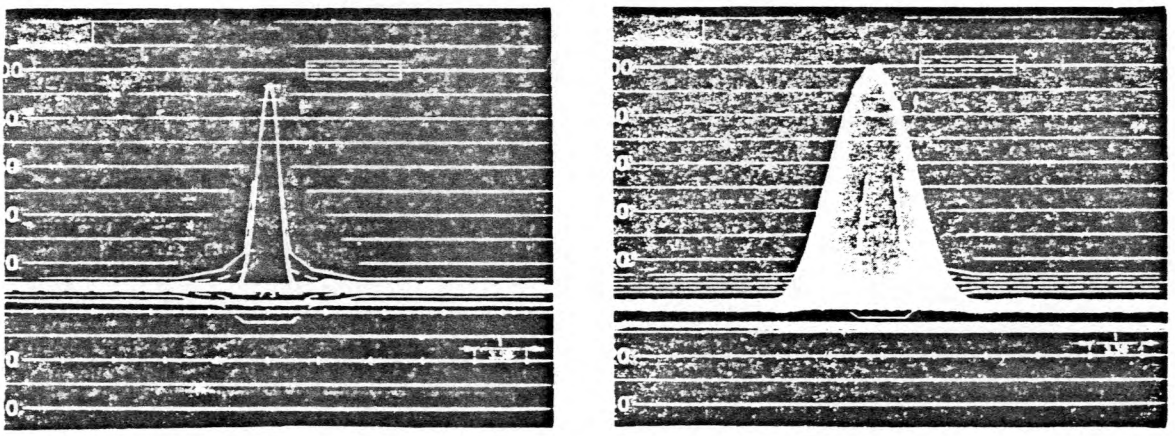
Modulator—Demodulator



Modulator – Processor (with delay equalization) – Demodulator

**Figure 10. Multiburst Response of Modulator — Demodulator and Modulator—Processor (With Delay Equalization) —Demodulator.**

achieve a flat delay from band edge to band edge would be tremendous. Such an equalizer would prove to be prohibitive not only from an alignment standpoint but also from a physical size standpoint, not to mention the problem of amplitude equalizing the device. Second, and even more important, the additional degree of improvement that would be obtained by complete equalization is negligible, compared to the improvement already achieved. This can be seen from Figure 11. Here we see the results of sending the Modulated 20T and $Sin^2$ 2T pulse test signals through the mod-processor (with equalizer)—demod. Note that the degree of distortion to the base line of the 20T pulse is virtually nonexistent, indicating, as is already evident from the swept delay curve, that the differential envelope delay at the video carrier frequency and color subcarrier frequency is zero. Even more important however, is the complete lack of undershoot and trailing smear that was evident when there was no processor equalization. This is due to the flat delay in the vicinity of the video carrier. Comparing Figure 10 to Figure 7, it is apparent the processor with equalization is transparent to the test signals discussed above.



K factor
≈2%
2T $Sin^2$ Pulse

Envelope Delay
≈0
Modulated 20T Pulse

**Figure 11. Video Test Signal Response of Modulator—Processor (With Delay Equalization) — Demodulator.**

Now that we have an equalized processor, with 60 dB of adjacent channel rejection, does not insure that the use of external traps and filters can be eliminated. Discussed earlier in this article is the fact that at the hub site many processors will have as inputs the desired signal plus many more, including equal level adjacents. To insure that no external traps/or filters are required at the input of each processor, it must have the capability of "handling" these signals without generating in-band intermodulation. This distortion could occur in any active stages prior to removal of the adjacent channels and once generated, no amount of adjacent channel rejection can remove it. To demonstrate this problem, I combined the output of a channel 2, channel 3, and channel 4 modulator. The carriers are not modulated to assist in looking at in-band intermodulation. The combined spectrum of these three modulators is shown in Figure 12.

These signals were then fed through a 6150 channel 3—3 processor with the input levels set first at +10 dBmV, then at +20 dBmV, and finally at +30 dBmV. Figure 13 is the output spectrum of the processor for these three levels. Note that with input levels of +30 dBmV, an in-band spurious signal is visible 1.5 MHz above the desired video carrier. Fortunately, this undesired signal is highly dependent upon the input levels and when they are set to +10 dBmV it is well below the noise level. It is a well known fact that due to the delayed AGC very little improvement is achieved in the output signal to noise ratio by feeding the processor with more than a +10 dBmV. Because of this no loss in output signal to noise ratio must be sacrificed in order to achieve sufficient "linearity" prior to removal of the adjacent channels.

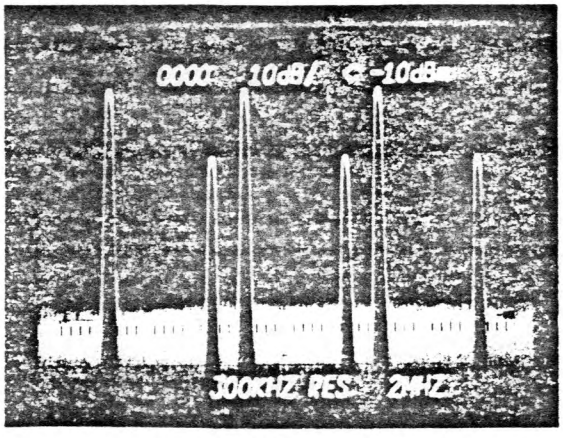
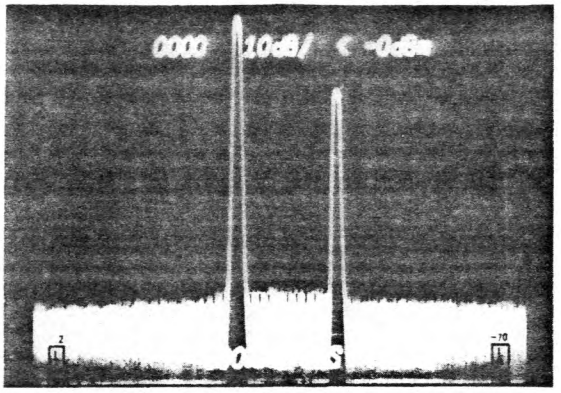
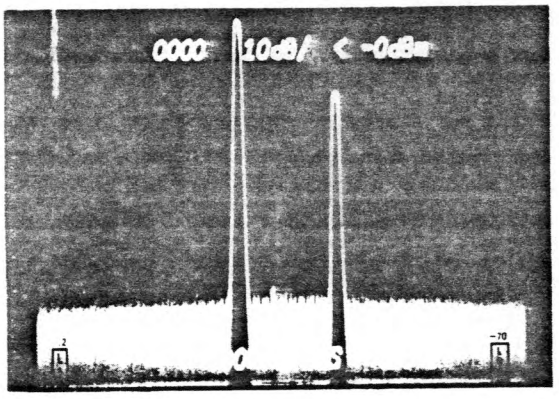


**Figure 12. Combined Output of Ch 2, Ch 3, and Ch 4 Modulator.**

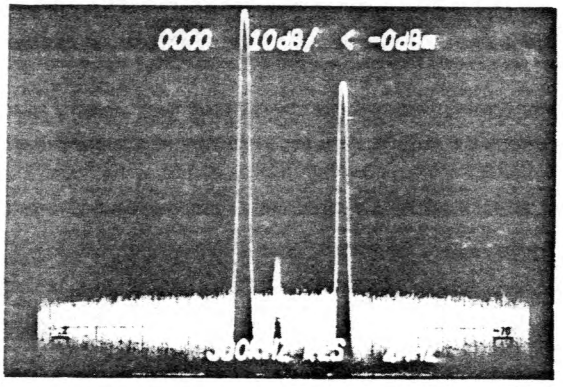
This article has discussed three important requirements for signal processors if they are to function successfully in modern day headend-hub systems. Certainly 60 dB adjacent channel rejection and the capability of handling equal level adjacent channels are a necessity if the processors at the hub site are to operate without external traps and/or filters. Delay equalization, if not a necessity is certainly desirable, especially in cases where the signal is processed twice. This by no means exhausts the list of desirable and in many cases necessary features for signal processors. Obviously, spurious free outputs from 5 MHz to 300 MHz are required if external



Input Level
+10 dBmV



Input Level
+20 dBmV



Input Level
+30 dBmV

**Figure 13. Output Spectrum of 6150 Ch 3 — Ch 3 ProcessorWith Equal Level Input Signals on Ch 2, Ch 3, and Ch 4.**

filters are to be eliminated. IF switching options, battery power capability, standby carrier modulation capability, phase lock capability are all desirable features. The requirements I discussed were chosen because headend-hub systems emphasized these limitations in first generation solid state signal processors.

[1] Frederick E. Terman, Electronics and Radio Engineering, Fourth Edition, Sec. 11-3, pp. 379-381.

[2] Tektronix, Nomographs for Measuring Relative Delay and Amplitude Using Modulated 20T Pulse for 625/50 Standards.

[3] T. M. Gluyas, Jr., "Television Transmitter Considerations in Color Broadcasting." RCA Review, September 1954, p.p. 312-334.

# APPLICATION OF PHASE LOCKED MODULATORS AND PROCESSORS

**By James O. Farmer**

**Scientific-Atlanta, Inc.**

Recently much as been said about the advantages phase locking techniques offer the CATV operator. This paper is intended to further acquaint the reader with the phase lock technique, so he can better utilize the available phase-locked equipment.

Phase lock is not new to the electronics industry. The first practical applications were coherent detectors in homodyne radio receivers built in the 1930's. However, the circuit complexity of a phase-locked loop is so great that until recently the technique was restricted to a few specialized applications. Now, thanks to progress in components and circuit techniques, phase lock is becoming economical for a wide range of applications.

CATV application of phase lock includes precision demodulation, minimizing the subjective effects of strong local TV signals, and various phase-locked headend schemes. This paper is concerned with only one of these applications of phase lock: minimizing the subjective effects of a strong local signal leaking into a subscriber's television set. This leakage problem has required the cable system to distribute strong signals on channels which are not locally used for TV broadcast. This off-channel conversion for cable distribution eliminates the leakage problem at the expense of available cable channels. New FCC rules and increasing program availability to cable systems have created pressure to use these fallow channels for some type of locally originated program. If the cable channel carrier is not exactly at the frequency of the interfering carrier, a co-channel beat will be seen on the subscriber's set.

Figure 1 illustrates the situation where two video carriers close together in frequency are present simultaneously. The signal on the left is the desired carrier, perhaps carrying a weather-scan

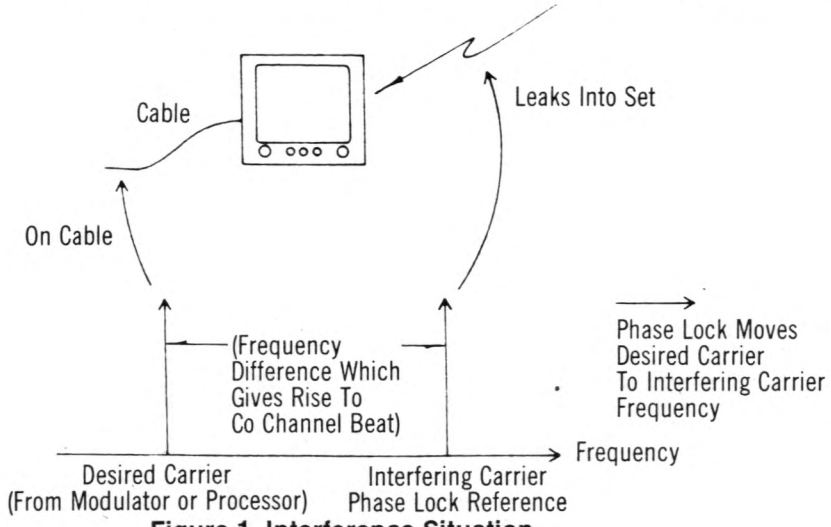



**Figure 1. Interference Situation**

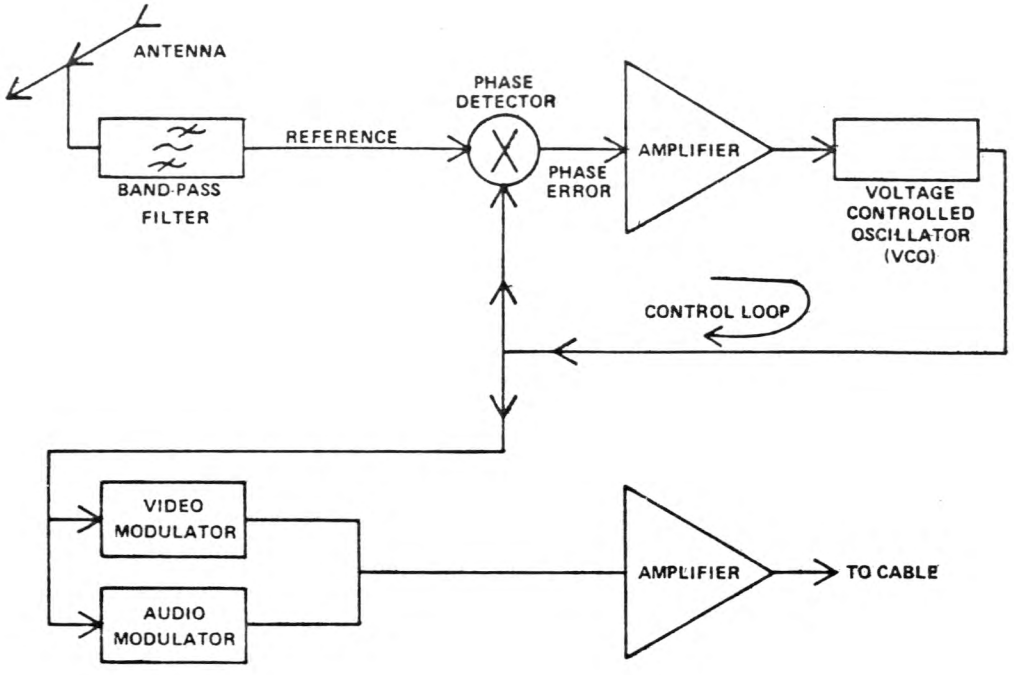



**Figure 2. Simplified Block Diagram, Phase-Locked Modulator**

picture. To the right is an interfering carrier which has leaked into the subscriber's TV set. When the two signals are detected. their frequency difference will create a component $\triangle f$ in the video. This shows up as black and white co-channel lines in the picture. The severity of the co-channel beat is dependent upon the relative levels of the two signals, and upon their frequency separation, $\triangle f$. Phase lock can be used here to force the desired carrier to the frequency of the interfering carrier, reducing $\triangle f$ to zero and eliminating the co-channel beat.

How do we accomplish this ? Figure 2 is a simplified block diagram of a phase-locked modulator. A sample of the interfering signal is picked off the air and applied to a fairly narrow bandpass filter. This filter removes all energy except the picture carrier and some of the luminance sidebands. The remaining signal is the reference, which is applied to a phase detector. The output of a voltage controlled oscillator is also applied to the phase detector. The output of the phase detector is a dc potential proportional to the cosine (in this case) of the phase difference between the reference signal and the voltage controlled oscillator (VCO) output. This phase error is applied to a loop amplifier (normally configured as an integrator for low frequencies), whose output is the VCO control voltage. Since the amplifier has very high gain at low frequencies, it will act on the VCO in whatever manner is necessary to reduce the phase error to nearly zero. Thus, the VCO output is required to maintain a strict phase relationship to the reference signal, and hence must operate at the same frequency. The VCO output is then applied to the modulator, and the remaining modulator operation is identical to that of a non phase-locked unit. In the case of Scientific-Atlanta 's Model 6300 PL Modulator, the modulation is performed at the 45.75 MHz intermediate frequency, which is then upconverted to the desired channel. A common local oscillator is used for up conversion and for down conversion of the reference. This assures phase coherency. For stability the VCO used is a voltage controlled crystal oscilator (VCXO).

Use of the reference signal directly instead of locking another oscillator to its frequency is not practical. This is because a bandpass filter cannot be realized that will adequately strip all modulation sidebands off the reference, maintaining this characteristic over all combinations of reference drift, time, and temperature. In this respect, the phase-locked loop operates as a tracking filter, since its output is at the frequency of the input, and all sidebands can be suppressed to an arbitary degree. Even so, the output can track the input carrier over a greater range than the effective filter bandwidth.

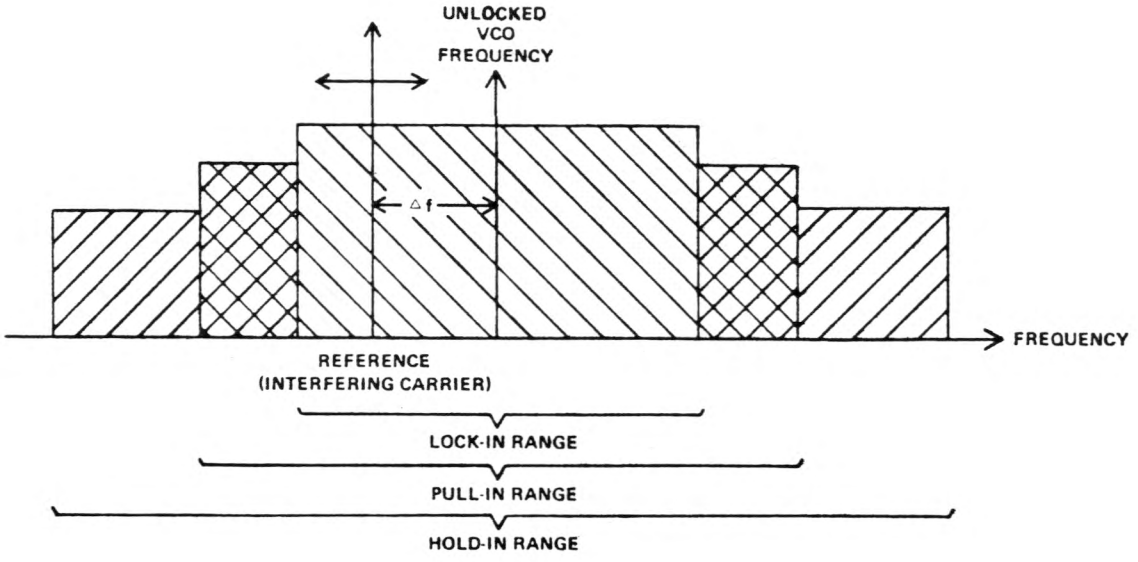



**Figure 3. Definition of Terms**

Before a phase-locked situation can exist, the VCXO must be made to run at the same frequency as the reference. Figure 3 defines some terms that are appropriate to the acquisition of phase lock. This figure is a spectrogram centered on the free running frequency of the VCXO; as this frequency changes, it will appear to stay fixed, but other features of the spectrogram will move. The free running frequency is the frequency at which the VCXO would run if no control voltage is applied from the loop amplifier. When power is first applied, or when the reference carrier first becomes available, the loop must somehow move the VCXO frequency to the reference frequency before phase lock can be achieved. Also shown in Figure 3 is the frequency of the reference, which might appear anywhere in the spectrogram.

The detailed process of bringing the VCXO into lock is quite complex and beyond the scope of this paper. However, a brief discussion of the process involved is in order. The difference between the reference and free running VCXO frequencies is defined as $\Delta f$. If the frequency difference is sufficiently small, the control loop is able to rapidly bring the VCXO into lock. The range over which this can be accomplished is called the lock-in range. For all practical purposes, the lock-in range is roughly equal to the loop bandwidth, which should be small in order to improve rejection of the reference sidebands. If the initial frequency difference is within the

lock-in range, the loop will rapidly acquire. If the initial frequency difference is out of the lock-in range but within the pull-in range, the loop can theoretically lock if given enough time. However, in this range practical considerations of amplifier offset and noise are such that lock generally cannot be achieved without additional circuitry. Outside of the pull-in range, lock cannot even theoretically be achieved without additional acquisition circuitry. However, once lock is achieved, the loop is able to hold lock over a much wider range, known as the hold-in range. This is the frequency difference over which, after lock is achieved, $\Delta f$ may drift and still permit the loop to remain locked. For the most useful type loop, the hold-in range is not a function of loop dynamics, but rather is a function of saturation levels of loop amplifiers or other components in the loop. Within certain practical bounds, the hold-in range may be made arbitrarily large.

Proper acquisition circuitry may be used to aid the loop in locking up at any frequency in the hold-in range. An adequate acquisition (and hold-in) range is one which permits acquisition under the worst possible initial frequency error, $\Delta f$, due to all causes. This acquisition must be without operator intervention. In order to determine the required acquisition range, we will develop a worst case error budget, taking into account all known sources of frequency error. Table 1 lists the major contributors to the error budget. The first source of error is the 1 kHz frequency tolerance imposed upon the broadcast station. Although not a frequency error, we will add in the 10 kHz broadcast carrier offset. If the acquisition range is insufficient to handle the carrier offset, then the offset must be specified when ordering the phase lock equipment or when transferring it to a different location. To the above tolerances must be added the drift of the VCXO, which might reach 5 kHz at the higher channels. This may be drift reduced to 0.2 kHz with a highly stable VCXO enclosed entirely in an oven. However, an oven is undesirable because the higher temperature will accelerate component failure, and will also increase power consumption. In addition, an oven stabilized oscillator must be allowed to warm up after turn-on or a power failure.

Another error which must be taken into account to insure that the operator does not have to "tweak" the phase lock after installation, is the initial frequency setting of the VCXO. A reasonable allowance for this is 0.5 kHz.

**TABLE 1. ERROR BUDGET FOR DETERMINATION OF REQUIRED ACQUISTION RANGE**

| | |
|---|---|
| **1. BROADCAST CARRIER TOLERANCE** | **1 kHz** |
| **2. BROADCAST CARRIER OFFSET** | **10** |
| **3. VCXO DRIFT (Ch. 13, non oven controlled)** | **5** |
| **4. VCXO INITIAL SETTING** | **.5** |
| **5. CRYSTAL AGING** | **:** |
| **6. COMPONENT AGING** | **?** |
| **7. VARIATION WITH POWER SUPPLY** | **?** |
| **ERROR BUDGET** | **16.5 kHz** |

**(For heterodyne processors, add down-conversion error)**

So far the error budget is ±16.5 kHz and there are still several other sources of error to consider. These errors are more difficult to quantitize. They include crystal aging, frequency fluctuation with power variations, etc. In the case of a phase locked heterodyne processor, we must also add the tolerances pertaining to the station whose signal we are processing. A safe acquisition range to consider is plus or minus 25 kHz. This allows for known sources of error, plus some margin for the undefined errors.

Implicit in the discussion of safe acquistion range has been the requirement that the loop must also exhibit an adequate hold-in range. Considerations with respect to adequate hold-in range include the use of servo compensation that develops a very high gain at low frequencies, so that a significant phase error does not develop at the phase detector output. Also, the VCO must be capable of being controlled over an adequate range. For good stability, the VCO should be crystal controlled, but special design techniques are required to pull a VCXO over the frequency range required.

Several types of acquisition circuitry have been developed over the years. Probably the most common technique in communications applications is use of a triangular ramp which drives the VCO back and forth over the entire hold-in range, searching for a signal to lock to. When the VCO is driven to within lock-in range, the loop locks and the ramp is disabled. The technique works, but is relatively slow, because acquisition dynamics limit the maximum search rate. Also, if significant sidebands exist within the search range, the loop may attempt to lock to them rather than to the carrier.

Recently several types of digital phase detectors have become popular for frequency synthesis work. The most elegant is a sequential circuit, available in I.C. form, that matches the negative-going transitions of the two waveforms. This detector has the property that if one input is higher in frequency than the other, the phase detector output is maximum in the direction that drives the VCXO into lock. Thus, if this detector is used, acquisition circuitry is unnecessary. For noise free applications such as frequency synthesis, the detector is often an excellent choice. However, it tends to be overly sensitive to noise or modulation sidebands, because of its characteristic of responding only to a waveform transition. In addition, its maximum operating frequency is only about 10 MHz, so it is not useful for our present application. Another acquisition scheme involves use of a frequency discriminator, which must be accurately zero'd with the VCXO free running frequency.

Scientific-Atlanta has used an acquisition technique in our 6300 PL phase locked modulator and 6150 PL processor which to our knowledge is unique. Before the loop pulls into lock, a beat note exists at the phase detector output. The characteristics of the beat note are analyzed to determine the direction in which the VCXO must be driven in order to acquire lock. This technique permits lock to be achieved in a few milliseconds, with an initial frequency difference of fifty kilohertz. Circuit logic makes the technique insensitive to false lock on the reference sidebands as long as their amplitude is less than that of the carrier.

The visual effect of non phaselock is primarily the familiar co-channel beats, the subjective effect being dependent upon the frequency difference. When the two signals are phaselocked, a ghost of the interfering signal will still be seen on the screen. No satisfactory method exists for eliminating this ghost except to eliminate the signal leakage. However, several things may be done in order to minimize the subjective effect of the ghosting. The subjective effect depends

upon the relative magnitude of the interference, upon picture content, upon the relative phase of the two carriers, and upon the relative frame rates of the two pictures.

Some reduction of the effect of ghosting may be obtained, in the case of display of text only, by utilizing white lettering on a black background. This is possible because black will then occupy the larger area, and a ghost is less noticeable with black picture content than with white. Black level is less affected by the ghost because of its higher carrier amplitude. Also a subjective effect is apparently at work here.

Phase angle between the two carriers is important because the effects of modulation of the interfering carrier on the overall instantaneous carrier amplitude may be shown to be dependent upon the manner in which the two carriers add vectorially. Unfortunately, this characteristic cannot be put to use in minimizing interference, because the relative phase of the two carriers at the home receiver cannot be maintained overtime.

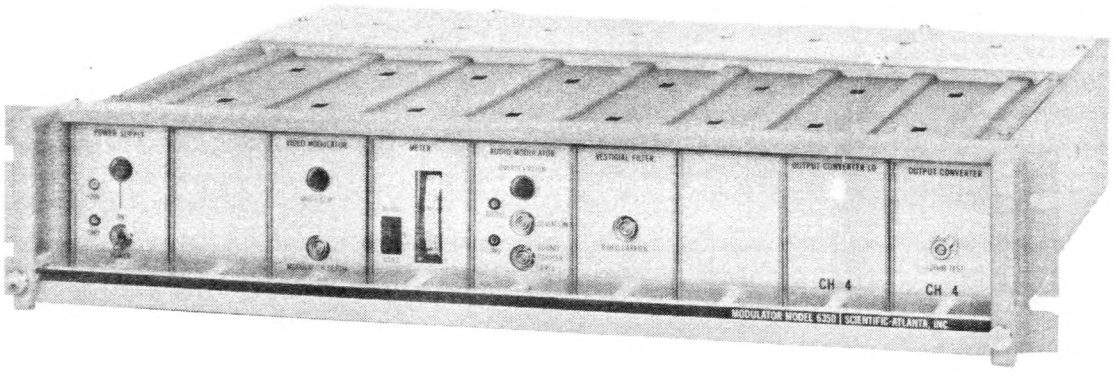
The relative frame rates of the two pictures are important because a moving ghost sync bar is much more objectionable than is a stationary bar. For this reason, the operator should be careful to operate at the same frame rate as that of the interfering signal, if maximum signal quality is to be maintained. This is normally less of a problem in all-color situations than in the situation of a black-and-white cable signal with line sync. The reason for this is that the field rate for color transmission is retarded slightly from the 60 Hz line rate. The effect of this retardation in the case of a 60 Hz field rate black-and-white cable picture being interfered with by a color picture, is to make the horizontal sync bar move through the picture in about .06 second. By comparison, if the local origination signal is also in color, and both broadcast and CATV sync generators meet FCC requirements of frequency stability, the sync bar will move across in a minimum of 11 seconds. Even though the moving sync bar is not much of a problem in all-color situations, the presence of two color subcarriers in the picture may, at times, cause a zoom effect, as if a zoom lens was being operated quickly. These problems may be eliminated if the local origination signal is genlocked to the interfering picture. (Note that genlock refers to locking the sync rates of the two pictures, and is independent of phaselock, which locks the frequency of the two RF carriers together.) This normally requires, in addition to a camera, a sync generator capable of being genlocked to external video. To provide this video source for genlocking purposes, Scientific-Atlanta has provided a utlity demodulator on its phaselocked modulator.

Phaselock techniques have several applications in CATV systems, but phaselock should not be looked upon as a "cure for what ails you." Intelligently applied, it can be useful in several applications, including elimination of co-channel beats. As with any other tool, it must be reasonably well understood before it can be used to good advantage. For this reason, the CATV engineer considering phaselock should be sure he understands the advantages and limitations of phaselock, in order to maximize his return on investment and avoid unnecessary disappointment.

Bibliography

1. Gardner, F.M., "Phaselock Techniques," Wiley, 1966.
2. More, A.W., "Phase-locked Loops for Motor Speed Control," IEEE Spectrum, Vol. 10, No. 4, April 1973, p. 61 f.f
3. Nash, G., "Phase-locked Loop Design Fundamentals," Application Note AN-535, Motorola Semiconductor Products, Inc., 1970.

# SECTION 4
# 6350 TELEVISION MODULATOR




Shown with optional meter.

**Figure 4-1. Model 6350 Television Modulator**

## Description

The Model 6350 Television Modulator generates a high quality vestigial sideband TV signal on all cable TV channels. Advanced circuit design results in the highest picture and sound fidelity from color or black and white video inputs and from either audio or 4.5 MHz subcarrier inputs. Scientific-Atlanta's pioneering application of the SAW* filter to CATV equipment results in unequalled vestigial sideband response characteristics. The SAW filter provides superior adjacent channel performance while retaining excellent group delay characteristics. Use of the SAW filter virtually eliminates vestigial sideband filter adjustments and provides excellent temperature stability.

A delay predistortion network is provided to conform with the standard FCC predistortion requirements. Front and rear panel test points are provided to assist in module trouble-shooting. Also provided are rear panel connectors for alternate IF inputs and outputs to facilitate program switching.

## Options

- The video switch detects loss of primary video and switches to alternate video input. (VS)
- Video AGC normalizes video level between different video sources such as two video tape recorders, or removes time dependent level changes from incoming video. (VAGC)
- Audio Switch option provides for selection of an alternate audio source, and also permits switching to emergency override audio. (AS)
- Audio Modulation Limiter prevents accidental overdeviation of the aural carrier (AML)
- IF Switch provides switching to an alternate IF source (HIFS)
- AGC control on alternate IF input provides for a constant output level when the alternate IF is being used. (IFAGC)
- Phase Lock of the output picture carrier is available. (PL)
- IF Loop Thru (IFL)
- Spectrum Inversion (SI)
- Meter (MM)
- Standby Battery Powering provides backup for failure 115V ac. (SP)
- Negative dc powering permits operations from a −19 to −31V dc battery supply in lieu of 115V ac. (DC)

*Surface acoustical wave.

# Model 6350 Television Modulator Specifications

**VIDEO INPUT**

Input Type
Composite video, sync. negative
Input Level
0.5v p-p minimum for 87.5% depth of modulation
Input Impedance
75 ohm, unbalanced (return loss 30 dB min) or high impedance loop-thru
Frequency Response
±0.5 dB from 10 Hz to 4.2 MHz
Differential Gain
±0.18 dB maximum at 87.5% modulation
Differential Phase
±0.5° maximum at 87.5% modulation
Hum and Noise
60 dB down with respect to 90% modulation
Tilt
1% maximum on 60 Hz, 50% square wave

**AUDIO INPUT**

Input Level
Normally set for −10 to +10 dBm
Internally switchable to two other ranges:
+5 to +25 dBm
−25 to −5 dBm
Input Impedance
600 ohm balanced. May be field modified for high impedance bridging input
Common Mode Hum Rejection
40 dB
Frequency Response
±0.5 dB from 30 Hz to 15 kHz
Carrier Shift with Modulation
±100 Hz or less
Harmonic distortion
0.5% maximum, 30 Hz to 15 kHz at 25 kHz deviation
FM Hum and Noise
60 dB down with respect to 25 kHz deviation
Intercarrier Frequency Tolerance
Within ±500 Hz to being 4.5 MHz above video carrier
−7° to +50°C)

**4.5 MHz FM INPUT**

Input Level
10 mV RMS to 200 mV RMS
Input Impedance
75 ohm unbalanced (normally supplied as composite picture and aural subcarrier input)

**RF OUTPUT**

Any standard VHF channel; sub-low; mid-band; super-band. Transposed video-sound channels available with optional spectrum inverter. Incremental and harmonically related coherent outputs are available, as are non-standard offset frequencies
Output Level
Output Impedance
75 ohm unbalanced (VSWR 1.35:1)
Output Level
+40 dBmV to +60 dBmV continuously variable
Spurious Outputs
60 dB below video carrier with video carrier at +60 dBmV and sound carrier at +45 dBmV
Frequency Tolerance Stability
±8 kHz VHF; mid-band
±10 kHz super-band
Group Delay Response
Meets FCC predistortion requirements for color transmission
Vestigial Sideband Response
−20 dB at channel edge
−40 dB at adjacent picture and sound carrier frequencies and all frequencies farther removed from channel

**PHASE LOCK OPTION**

Reference Input Frequency
Available in same frequency selections as output video carrier
Reference Input Level
−10 dBmV to +40 dBmV
Acquisition Range
±25 kHz
Demodulated Video Output
1V p-p nominal, negative sync. Can be used for genlocking to interfering modulation. Color burst included

**GENERAL**

Input Voltage
100 to 130V ac, 50 to 60 Hz
+24V dc nominal standby battery with optional built-in automatic switchover
−24V dc nominal battery operation with optional dc converter module
Input Power
40 watts
Weight
10.4 kg (23 pounds)
Mounting
89mm (3½")H x 483mm (19")W x 495mm (19½")D
(Standard rack mount spacing; rack slides optional)
Operating Temperature Range
−7 to +50°C (+20 to +120°F)

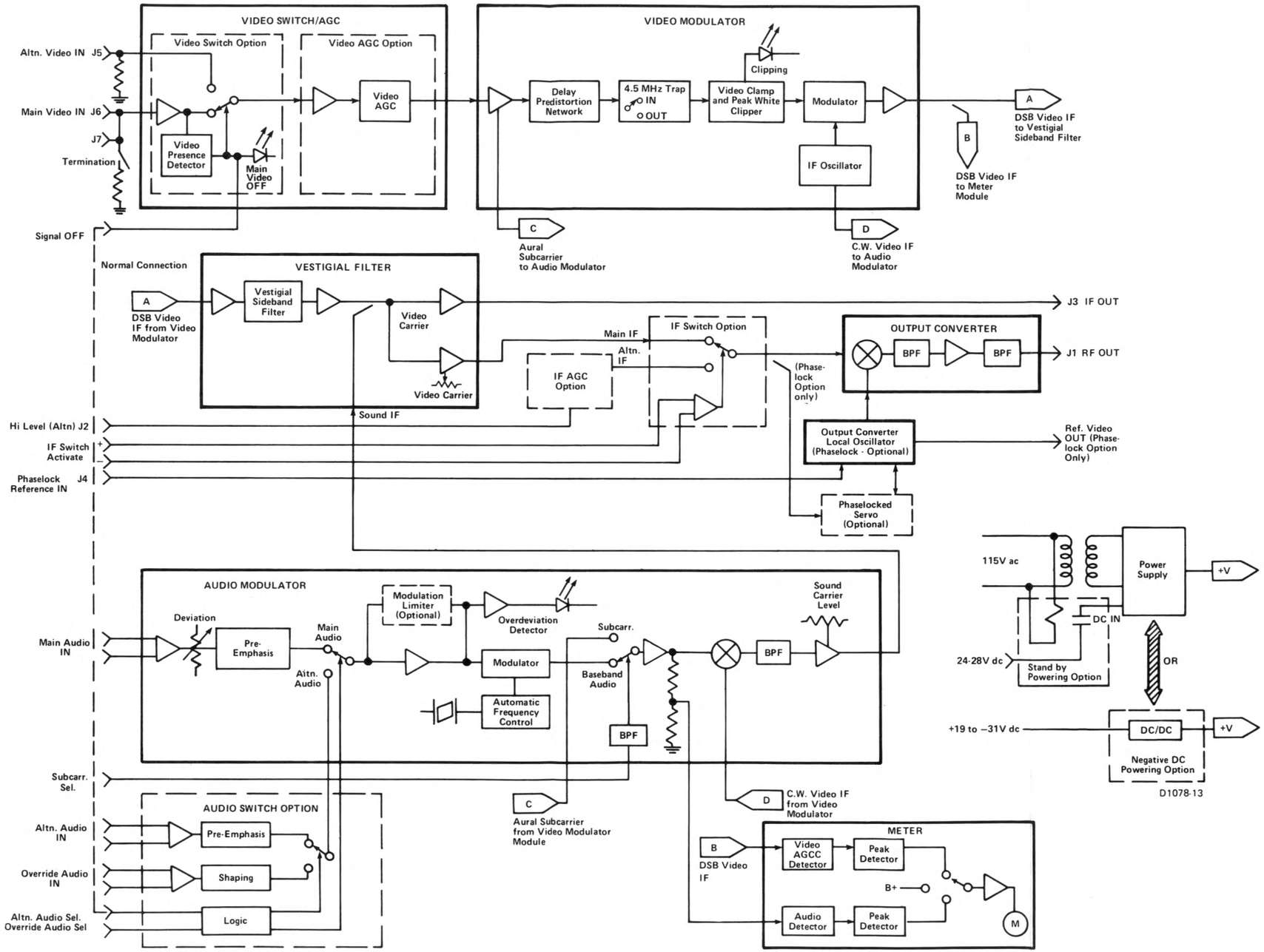
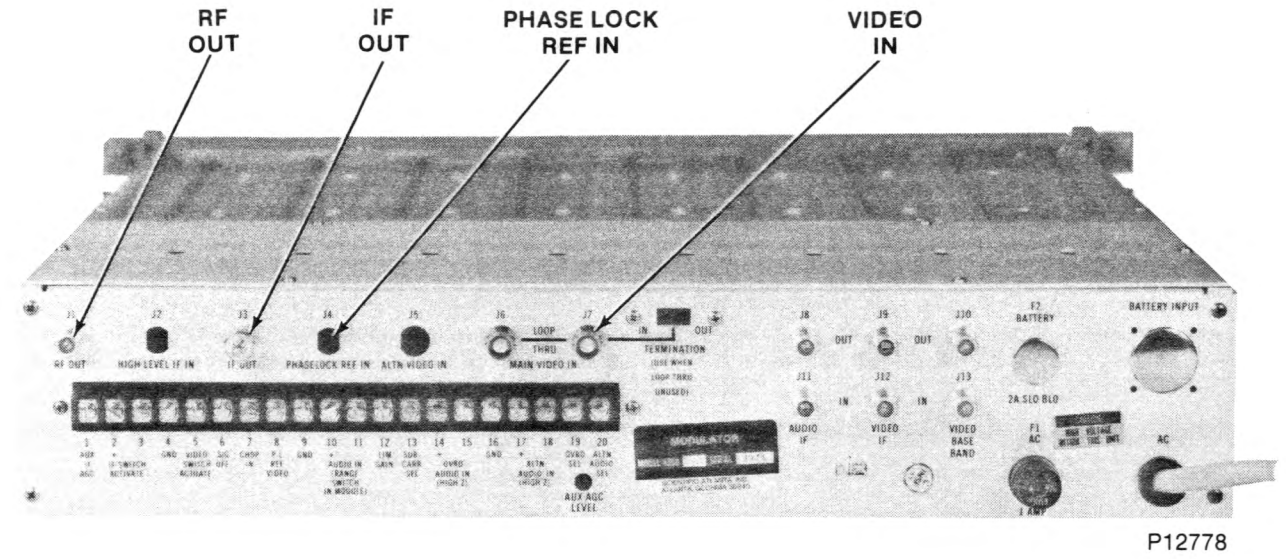



**Figure 4-2. Model 6350 Television Modulator Block Diagram**



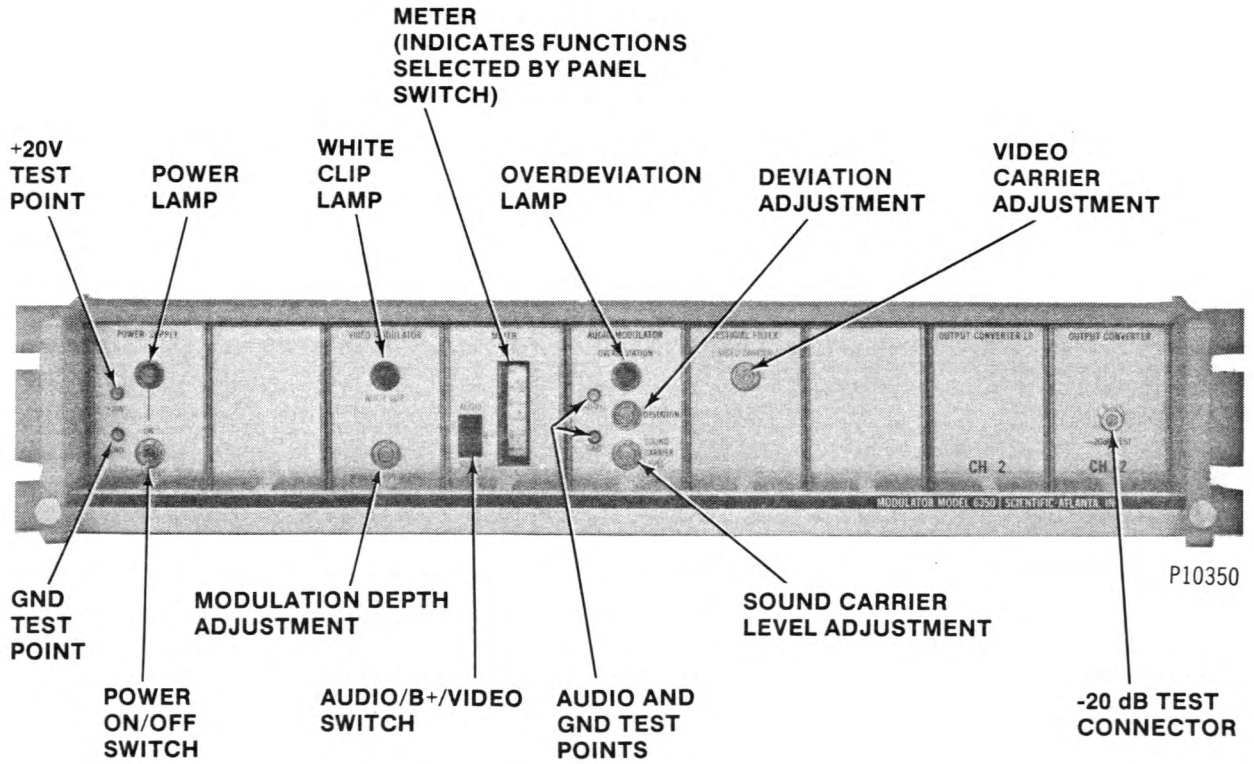




**Figure 4-3. Model 6350 Front and Rear Panel Views**

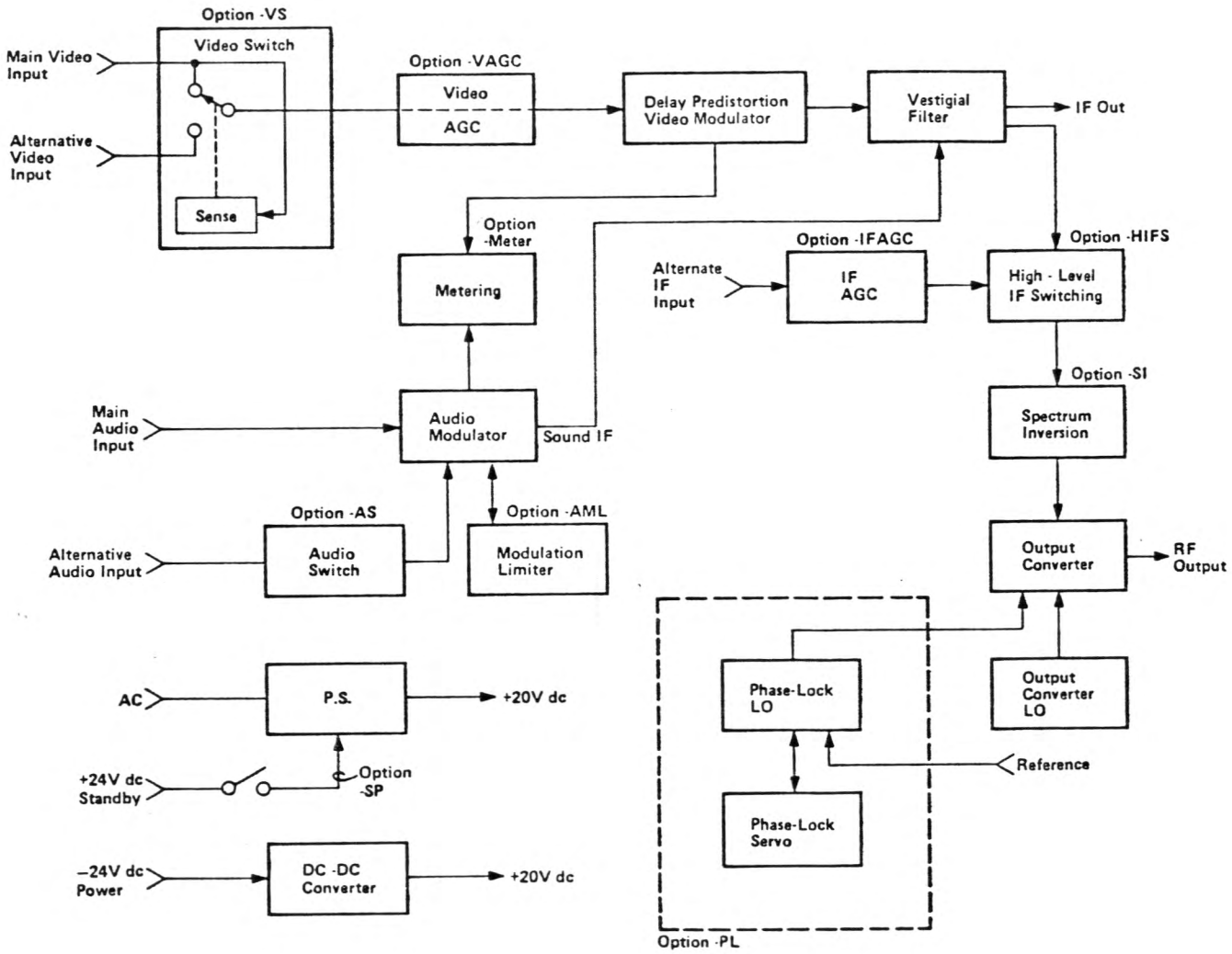



**Figure 4-4. Model 6350 Block Diagram**

### Sync Suppression Systems (SIFL Option)

Sync suppression systems used in older equipment require a composite video and audio IF after the vestigial sideband filter (VSF). The 6350 SIFL option provides these connections at the real panel as shown. The signals from the scrambler are returned to the output converter of the modulator.

### Pulse-Sync Suppression (DIFL - A Option)

Pulse-sync suppression systems such as the Scientific-Atlanta 6700 - or Jerrald (SSE 200AN) require separate video and audio IF signals prior to the SAW VSF. The 6350 DIFL - A option provides these connections at the rear panel as shown. Signals from the scrambler are returned to pass through the VSF.

### Sinewave Suppression (DIFL - B Option)

Sinewave suppression systems such as the Oak Mark V require separate video and audio IF signals after the VSF. The 6350 DIFL - B option provides these connections at the rear panel as shown. Signals from the scrambler are returned to the output converter.

### DC Sync Suppression Systems (ZSM Option)

DC sync suppression systems such as that used by Zenith Radio Corporation use the ZSM option to the 6350. Connections on the rear panel are for a subcarrier input and synchronizing input from the scrambler.

**NOTE**

**In addition to the ZSM Option, the Video Modulator Module must be modified by Zenith Radio Corporation.**

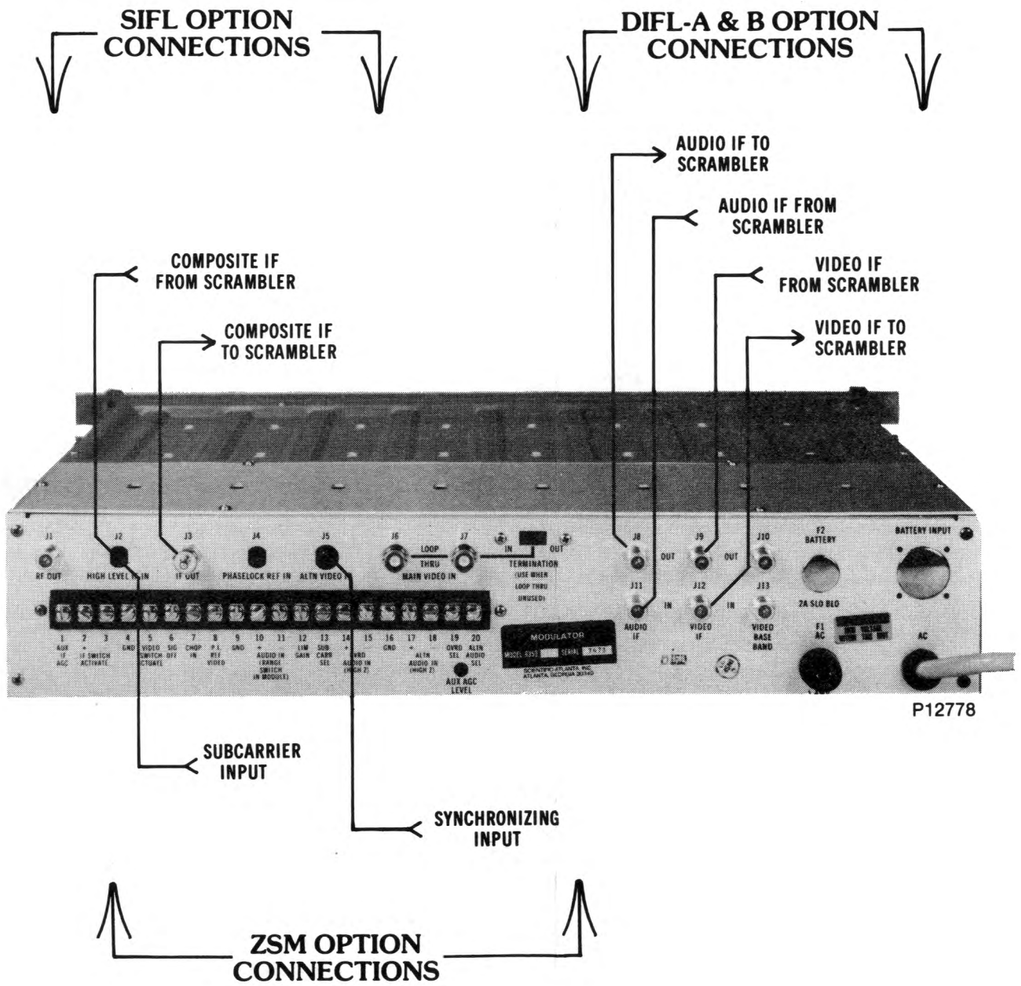
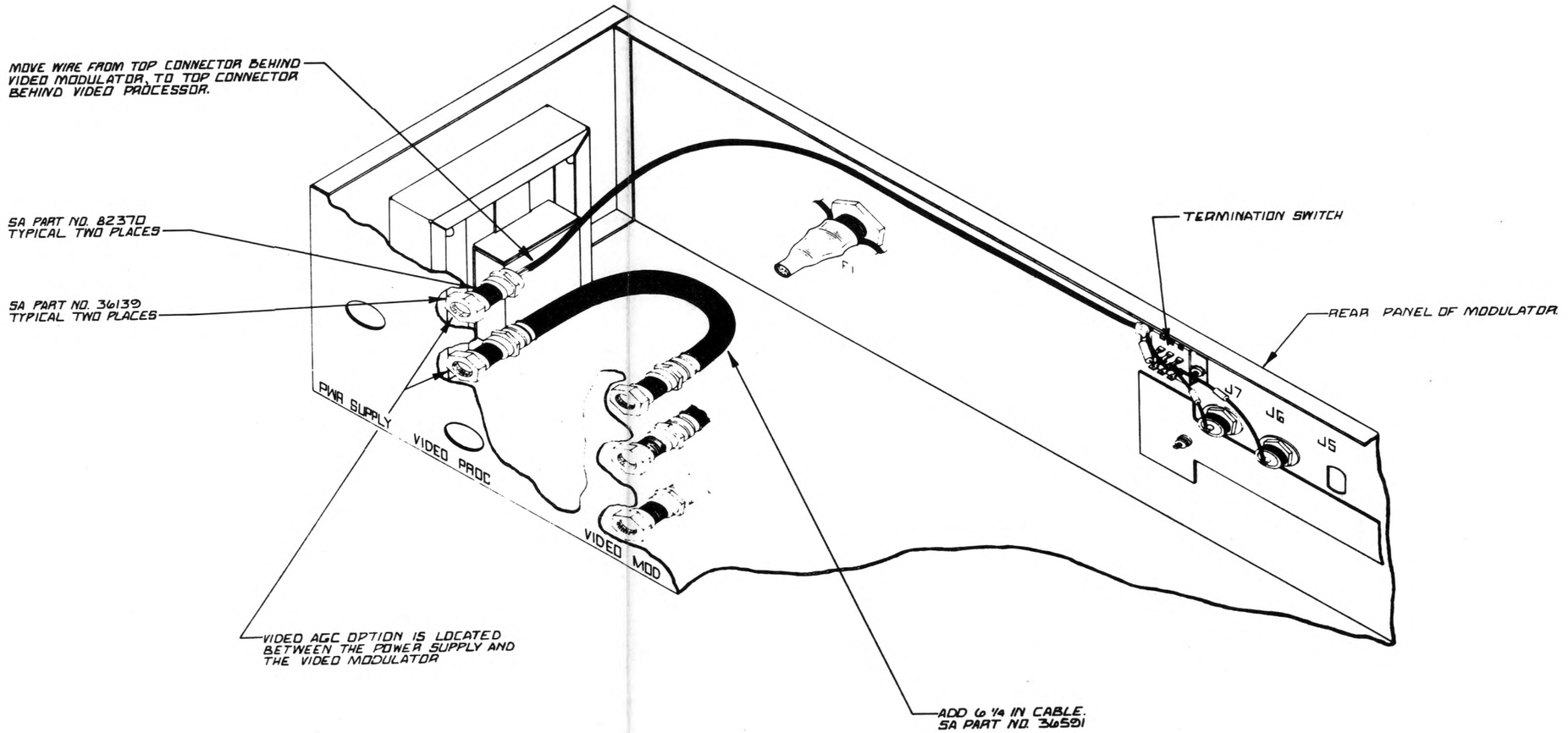


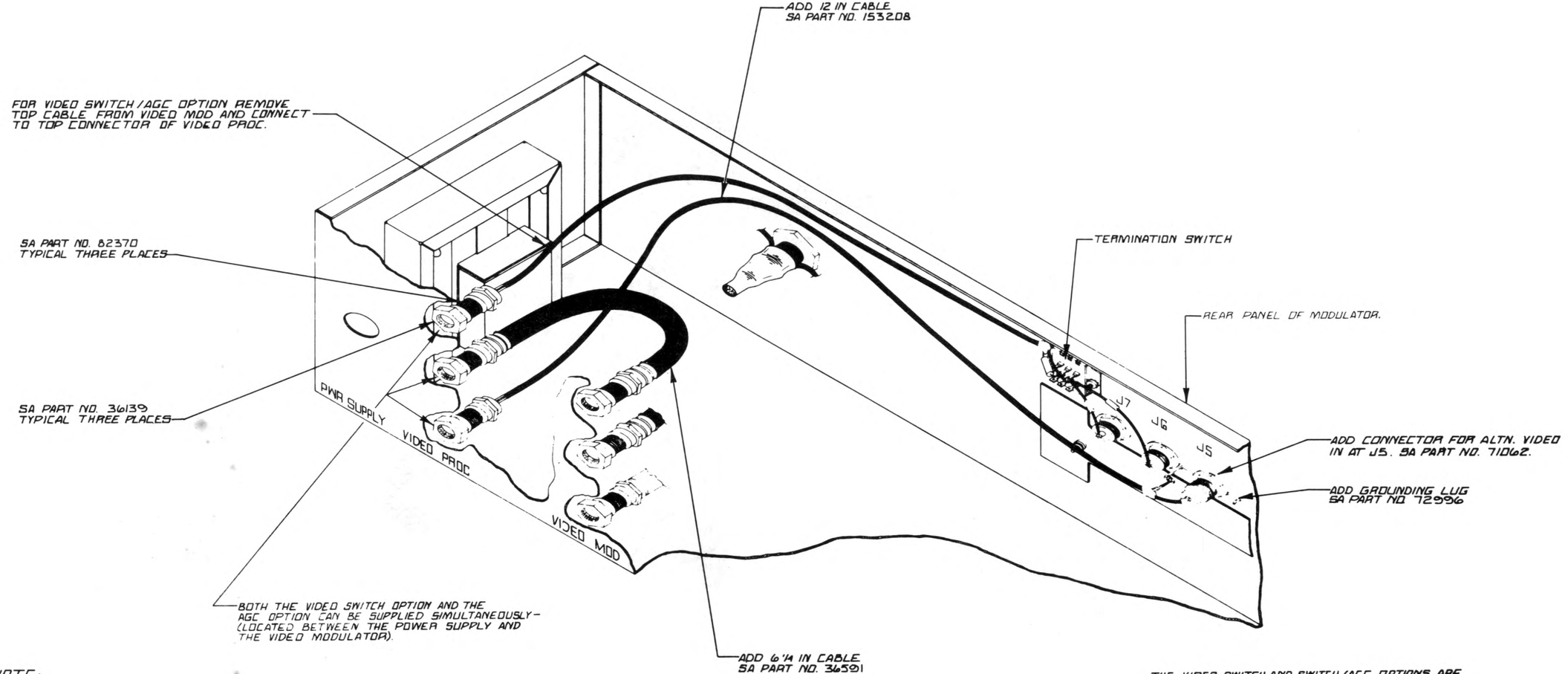

**Figure 4-5. Model 6350 Modulator/Scrambler Interface Options and Associated Rear-Panel Connections.**


MOVE WIRE FROM TOP CONNECTOR BEHIND VIDEO MODULATOR, TO TOP CONNECTOR BEHIND VIDEO PROCESSOR.
SA PART NO. 82370 TYPICAL TWO PLACES
SA PART NO. 36139 TYPICAL TWO PLACES
PWR SUPPLY
VIDEO PROC
VIDEO MOD
F1
TERMINATION SWITCH
REAR PANEL OF MODULATOR
J7
J6
J5
VIDEO AGC OPTION IS LOCATED BETWEEN THE POWER SUPPLY AND THE VIDEO MODULATOR
ADD 6 1/4 IN CABLE. SA PART NO. 36591

THE VIDEO AGC OPTION IS IN CONJUNCTION WITH STANDARD CABLING.

162617

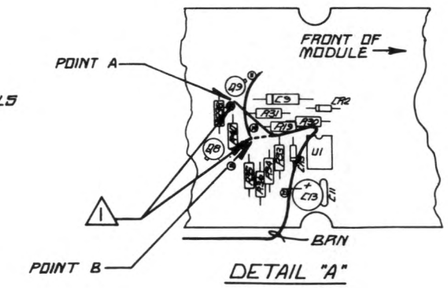
**Figure 4-6. Model 6350 Video AGC Installation Diagram**


ADD 12 IN CABLE
SA PART NO. 153208
FOR VIDEO SWITCH/AGC OPTION REMOVE TOP CABLE FROM VIDEO MOD AND CONNECT TO TOP CONNECTOR OF VIDEO PROC.
SA PART NO. 82370
TYPICAL THREE PLACES
SA PART NO. 36139
TYPICAL THREE PLACES
PWR SUPPLY
VIDEO PROC
VIDEO MOD
TERMINATION SWITCH
REAR PANEL OF MODULATOR.
J7
J6
J5
ADD CONNECTOR FOR ALTN. VIDEO IN AT J5. SA PART NO. 71062.
ADD GROUNDING LUG SA PART NO 72996
BOTH THE VIDEO SWITCH OPTION AND THE AGC OPTION CAN BE SUPPLIED SIMULTANEOUSLY - (LOCATED BETWEEN THE POWER SUPPLY AND THE VIDEO MODULATOR).
ADD 6'14 IN CABLE SA PART NO. 36501

THE VIDEO SWITCH AND SWITCH/AGC OPTIONS ARE IN CONJUNCTION WITH STANDARD CABLING.

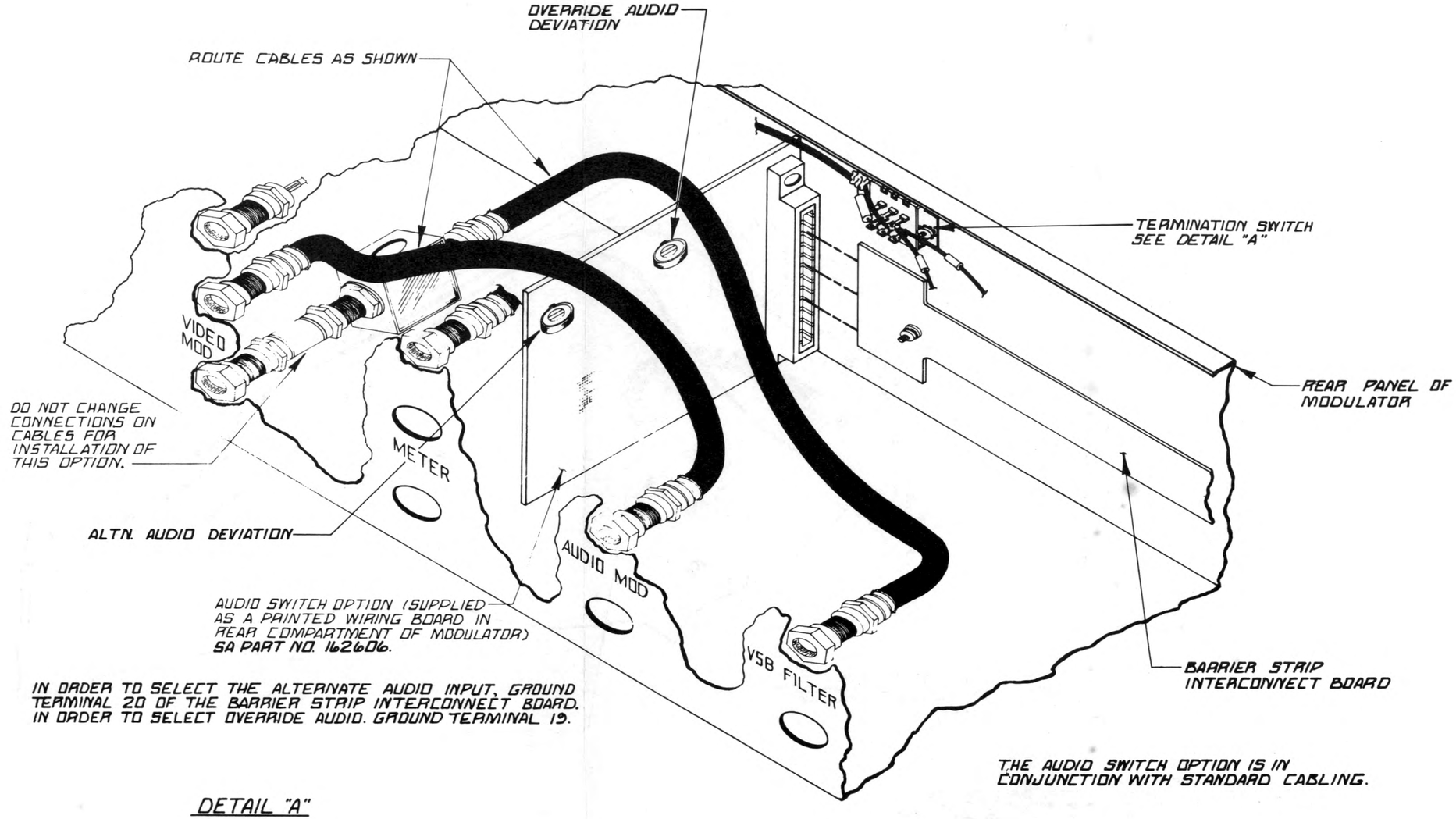
NOTE:

⚠ VIDEO SWITCH ONLY, DETAIL "A":
BRN WIRE SHOULD BE AT POINT A EXCEPT WHEN THE MAIN VIDEO SIGNAL IS ACCOMPANIED BY SUBCARRIER AUDIO AND THE ALTERNATE VIDEO IS ACCOMPANIED BY BASEBAND AUDIO. IN THIS CASE, MOVE WIRE TO POINT B. THE BASEBAND AUDIO SHOULD BY SUPPLIED TO TERMINALS 10 AND 11 ON THE REAR OF THE MODULATOR.
IF AN AUDIO SWITCH OPTION IS USED, STRAP TERMINALS 6 AND 20 ON THE REAR. IF ONE SOURCE UTILIZES BASEBAND AUDIO AND THE OTHER UTILIZES SUBCARRIER AUDIO, STRAP TERMINALS 6 AND 13.




162615

**Figure 4-7. Model 6350 Video Switch Installation Diagram**


OVERRIDE AUDIO DEVIATION
ROUTE CABLES AS SHOWN
TERMINATION SWITCH SEE DETAIL "A"
REAR PANEL OF MODULATOR
VIDEO MOD
METER
AUDIO MOD
VSB FILTER
DO NOT CHANGE CONNECTIONS ON CABLES FOR INSTALLATION OF THIS OPTION.
ALTN. AUDIO DEVIATION
AUDIO SWITCH OPTION (SUPPLIED AS A PAINTED WIRING BOARD IN REAR COMPARTMENT OF MODULATOR) SA PART NO. 162606.
BARRIER STRIP INTERCONNECT BOARD
THE AUDIO SWITCH OPTION IS IN CONJUNCTION WITH STANDARD CABLING.

IN ORDER TO SELECT THE ALTERNATE AUDIO INPUT, GROUND TERMINAL 20 OF THE BARRIER STRIP INTERCONNECT BOARD. IN ORDER TO SELECT OVERRIDE AUDIO. GROUND TERMINAL 19.

DETAIL "A"

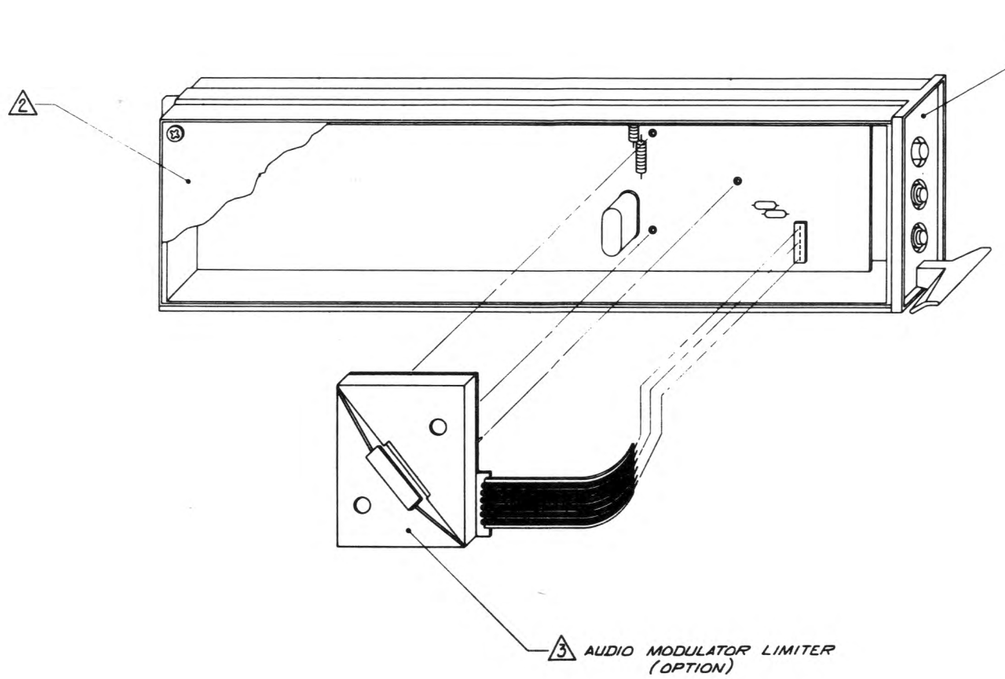
TERMINATION SWITCH

162618

NOTE:

1. IT MAY BE NECESSARY TO LOOSEN AND MOVE GROUNDING LUG FROM HORIZONAL POSITION TO VERTICAL POSITION IN ORDER TO ALLOW CLEARANCE FOR THE AUDIO SWITCH BOARD.

**Figure 4-8. Model 6350 Audio Switch Installation Diagram**

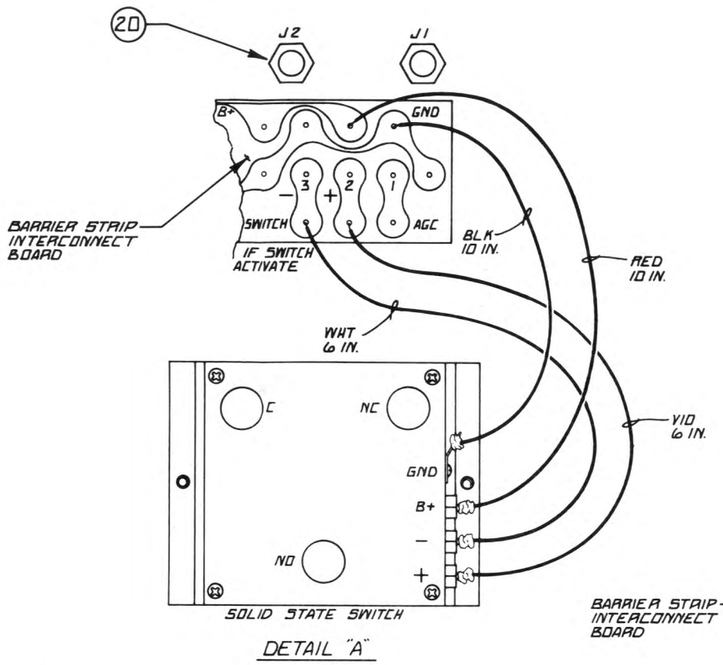



162619

*INSTALLATION PROCEDURE*

1. REMOVE AUDIO MODULATOR FROM MAIN CHASSIS.
2. REMOVE LEFT SIDE COVER FROM MODULE.
3. PLUG IN AUDIO MODULATOR LIMITER PRINTED WIRING BOARD OPTION AND RIBBON CONNECTOR AS SHOWN.

**Figure 4-9. Model 6350 Audio Modulation Limiter Installation Diagram**

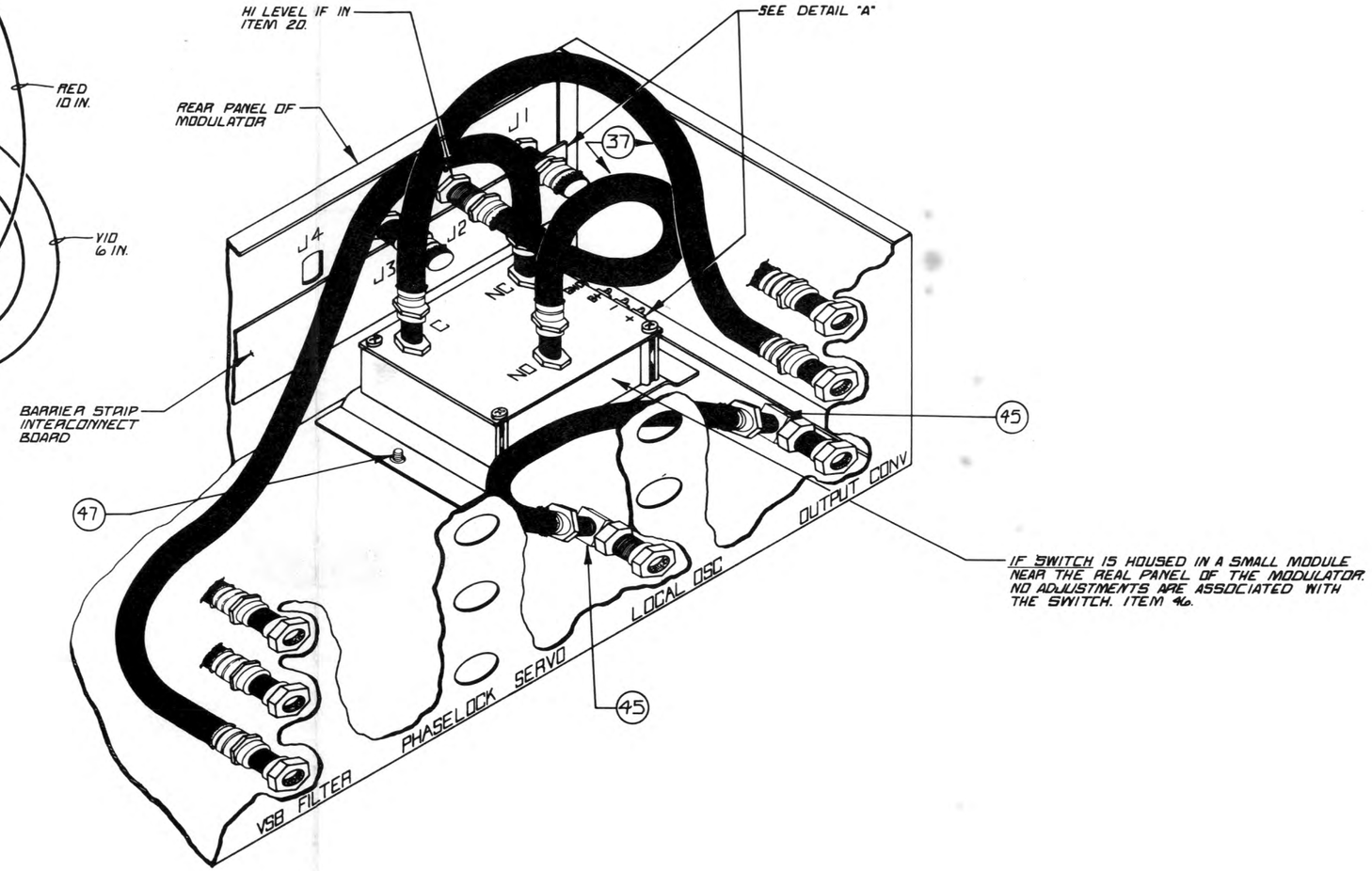



DETAIL "A"

IF SWITCH ACTIVATION:

1. GROUND TERMINAL 3 ON THE REAR PANEL BARRIER STRIP AS SHOWN, OR
2. APPLY A VOLTAGE IN EXCESS OF 5 VOLTS BUT LESS THAN 20 VOLTS TO TERMINAL 2 OF THE BARRIER STRIP.

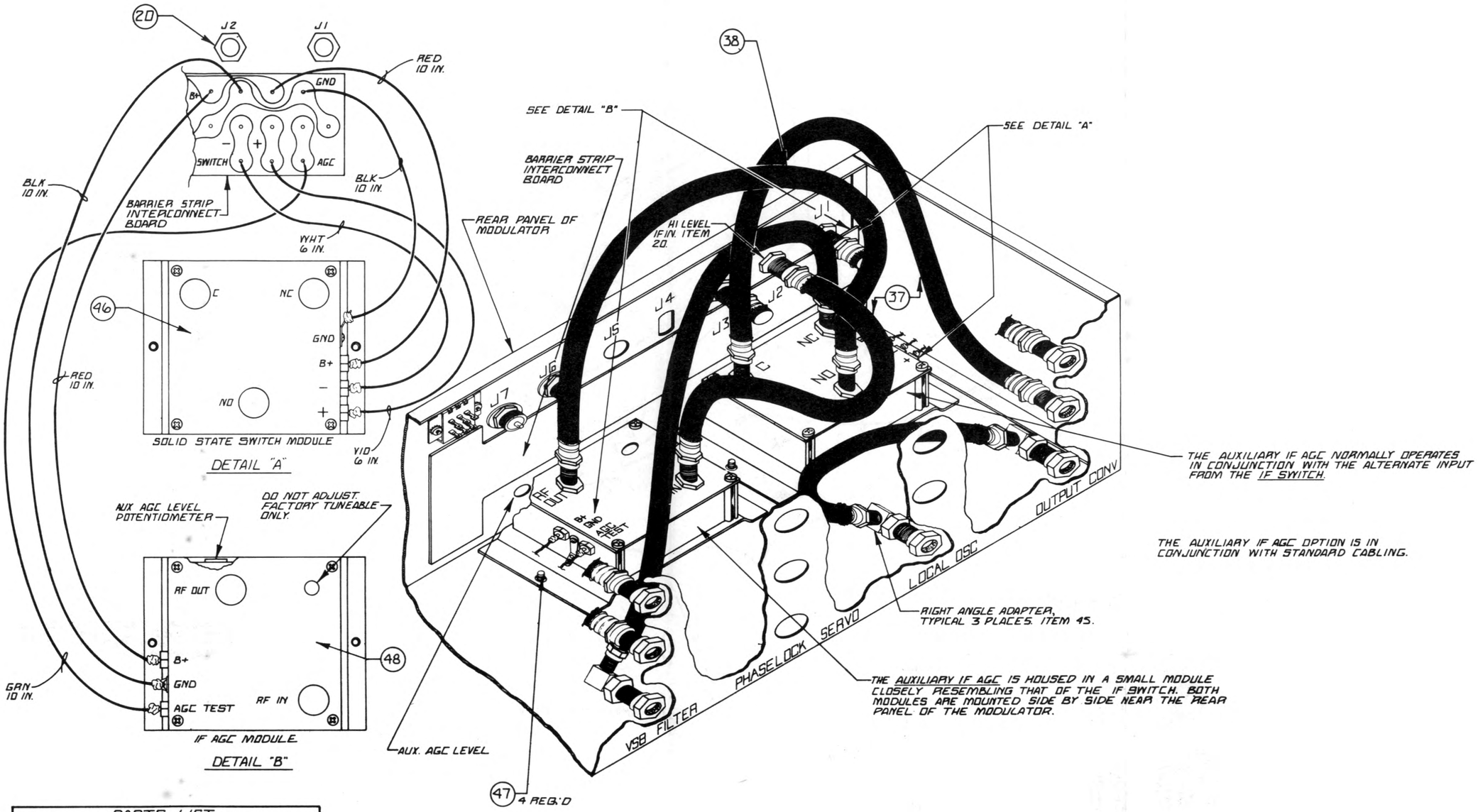
| PARTS LIST | | |
|---|---|---|
| ITEM NO. | DESCRIPTION | SA P/N |
| 20 | CONNECTOR "F" BULLET | 360200 |
| 37 | CABLE, 7 IN. | 360594 |
| 45 | RIGHT ANGLE ADAPTER (ELBOW) | 86826 |
| 46 | SOLID STATE SWITCH (MODULE) | 360595 |
| 47 | SCREW, 4-40 x 1/4 IN. | 72317 |
| | | |
| | | |




THE IF SWITCH OPTION IS IN CONJUNCTION WITH STANDARD CABLING.

**Figure 4-10. Model 6350 IF Switch Installation Diagram**

164235

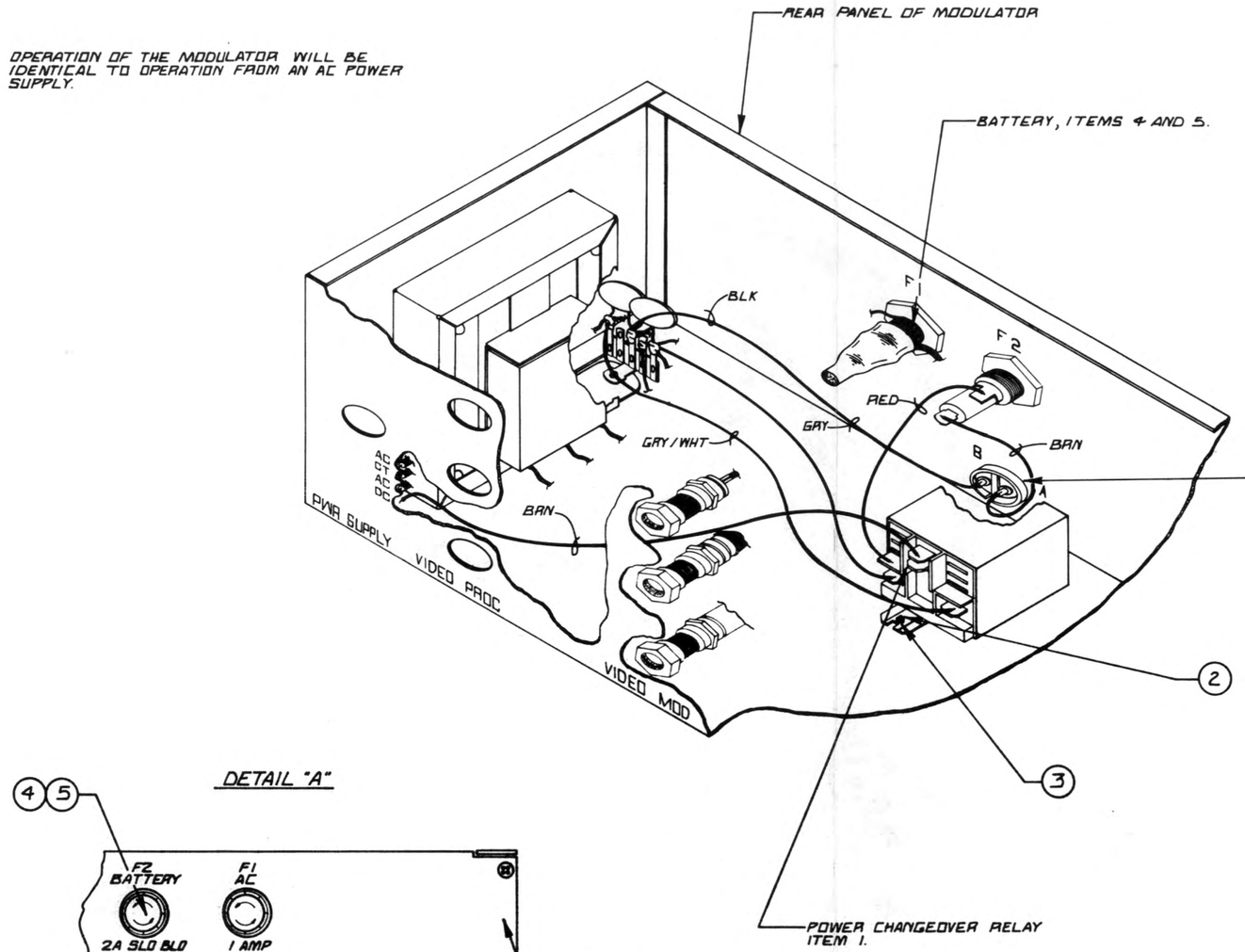



| PARTS LIST | | |
|---|---|---|
| ITEM NO. | DESCRIPTION | SA P/N |
| 20 | CONNECTOR, "F" BULLET | 36200 |
| 37 | CABLE, 7 IN. | 36594 |
| 38 | CABLE, 9 3/4 IN. | 36592 |
| 45 | RIGHT ANGLE ADAPTER (ELBOW) | 86826 |
| 46 | SOLID STATE SWITCH (MODULE) | 36595 |
| 47 | SCREW 4-40 x 1/4 | 72317 |
| 48 | IF AGC MODULE | 160610 |
| | | |
| | | |
| | | |

164236

**Figure 4-11. Model 6350 Auxiliary IF AGC Installation Diagram**

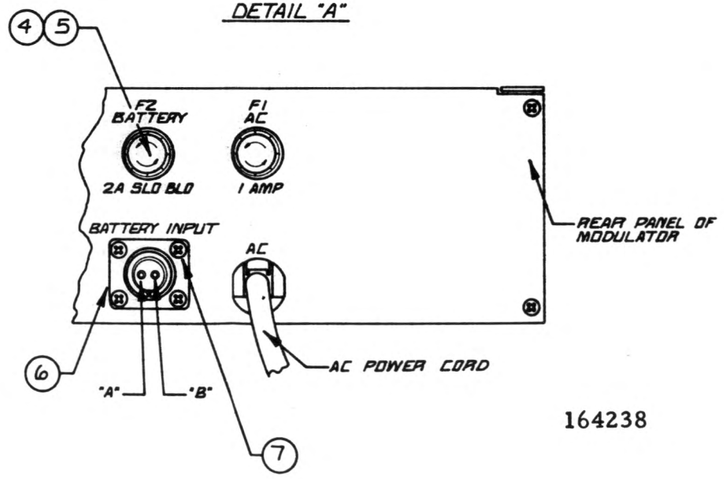
OPERATION OF THE MODULATOR WILL BE IDENTICAL TO OPERATION FROM AN AC POWER SUPPLY.




| PARTS LIST | | |
|---|---|---|
| ITEM NO. | DESCRIPTION | SA P/N |
| 1 | POWER CHANGEOVER RELAY | 82760 |
| 2 | SCREW, 4-40 X 5/16 | 72319 |
| 3 | NUT, FLEXLOCK | 71496 |
| 4 | FUSE HOLDER | 71249 |
| 5 | FUSE 2 AMP SLO BLO | 71224 |
| 6 | CONNECTOR 3102A 12S 3P | 70269 |
| 7 | SCREW, 4-4 1/4 | 82756 |
| | | |
| | | |

STANDARD CABLING IS IN CONJUNCTION WITH STANDBY +DC POWER OPTION. CONNECT BRN, RED, GRY, BLK, AND GRY/WHT WIRES AS INDICATED.

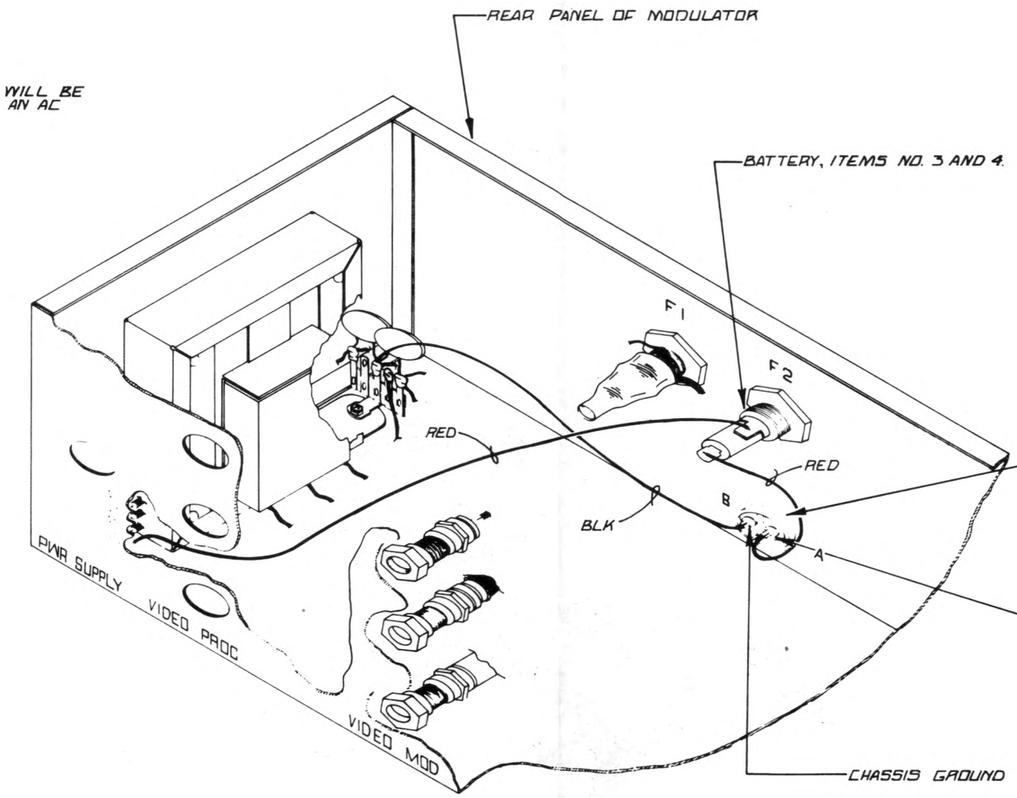
DETAIL "A"



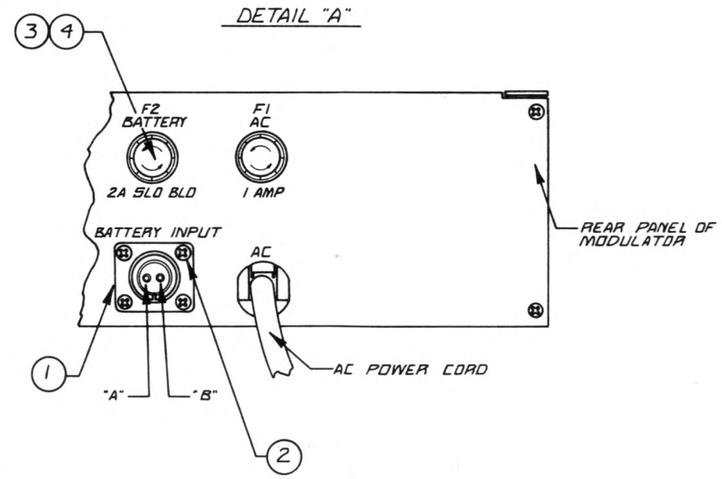

164238

**Figure 4-12. Model 6350 Standby +DC Powering Option Installation Diagram**

OPERATION OF THE MODULATOR WILL BE IDENTICIAL TO OPERATION FROM AN AC POWER SUPPLY.




| PARTS LIST | | |
|---|---|---|
| ITEM NO. | DESCRIPTION | SA P/N |
| 1 | CONNECTOR 3102A 12S 3P | 70869 |
| 2 | SCREW, 4 X 4-1/4 TYPE AB | 82756 |
| 3 | FUSE, 3 AMP SLO BLO | 71228 |
| 4 | FUSE HOLDER | 71249 |
| | | |
| | | |




164239

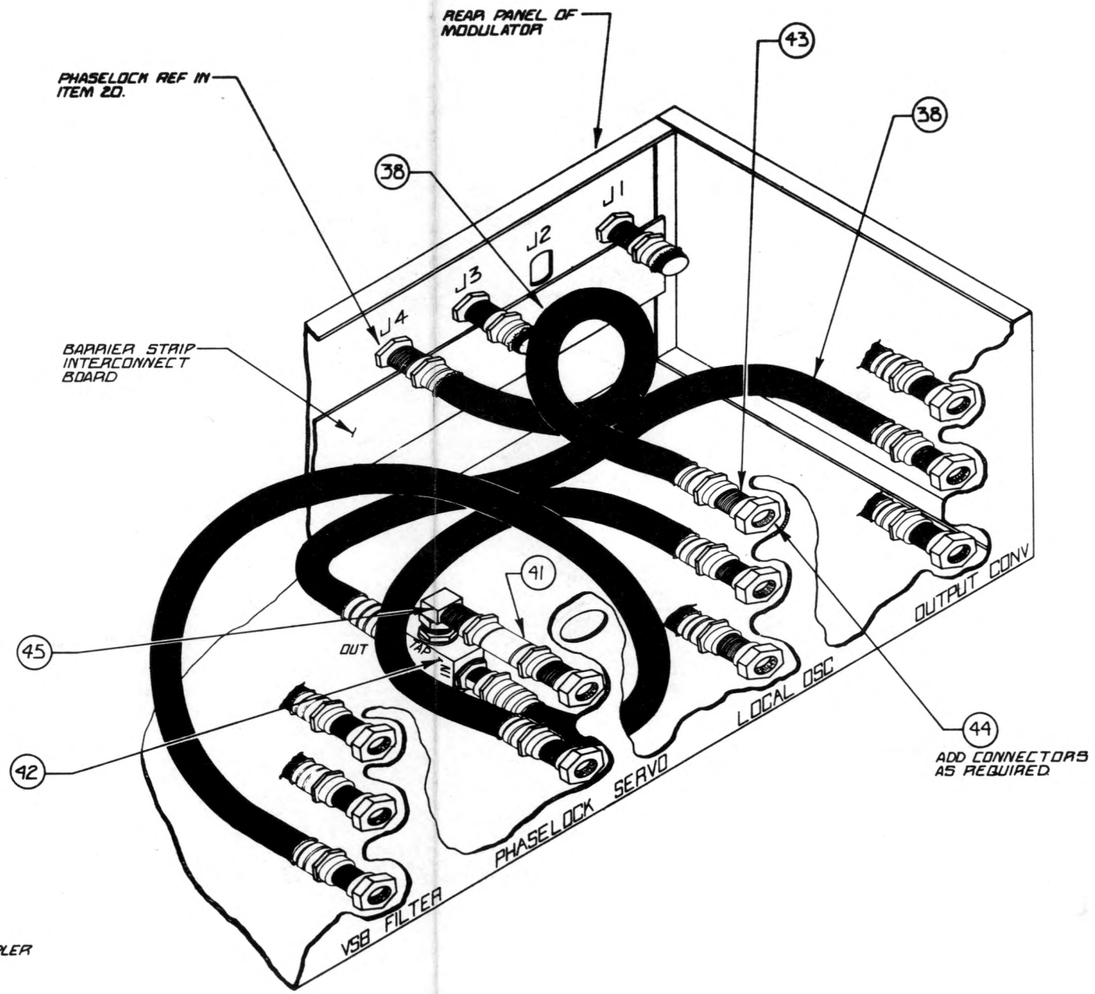
STANDARD CABLING IS IN CONJUNCTION WITH THE NEGATIVE DC POWER OPTION. CONNECT RED AND BLK WIRES AS INDICATED.

**Figure 4-13. Model 6350 Negative Battery Powering Option Installation Diagram**

| PARTS LIST | | |
|---|---|---|
| ITEM NO. | DESCRIPTION | SA P/N |
| 20 | CONNECTOR "F" BULLET | 36200 |
| 36 | CABLE, 6 1/4 IN. | 36591 |
| 38 | CABLE, 9 3/4 IN. | 36592 |
| 41 | CONNECTOR "F" MALE TO "F" MALE | 81516 |
| 42 | DIRECTIONAL COUPLER 10 DB | 82740 |
| 43 | CHASSIS CONNECTOR "F" FEM. TO MALE PUSH ON | 82370 |
| 44 | NUT | 36139 |
| 45 | RIGHT ANGLE ADAPTER (ELBOW) | 86826 |
| | | |

IN SOME CASES, ADDITIONAL OPTIONS MAY BE SUPPLIED, OR STYLES OF DIRECTIONAL COUPLERS USED, WHICH PROHIBIT INSTALLATION AS SHOWN. ALTERNATE PLACEMENT OF THE DIRECTIONAL COUPLER IS ACCEPTABLE, AS LONG AS WIRING IS CORRECT.

164237




PHASELOCK OPTION IS IN CONJUNCTION WITH STANDARD CABLING. NO ADJUSTMENTS NECESSARY.

**Figure 4-14. Model 6350 Phase Lock Option Installation Diagram**

# SETTING AND MAINTAINING MODULATION LEVELS

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## Introduction

We deal here with two related problems faced by the user of television modulators: the problem of initially setting modulation and the problem of maintaining it over a long period of time. Actually, we have four problems here, because the same problems exist for both audio and video. Failure to maintain correct video modulation may result in loss of sync, a washed out picture, and weak color in the case of undermodulation. Overmodulation can result in poor detector, amplifier, and kinescope linearity, causing problems with the intensity and tint of picture highlights, and can cause excessive sync buzz in the audio channel. Undermodulation in the audio channel results in weak audio and poor signal to noise ratio. Overmodulation of the audio can result in distortion. Any errors in setting modulation levels are particularly annoying when changing from one channel to another. Signal processors will generally not alter the modulation level of off-air stations, though frequency response errors will cause an apparent change in video modulation depth at higher video frequencies.

Let's first consider the problem of setting (i.e., measuring) the video modulation level, then we can consider the problem of maintaining it. After that, we will do the same for the audio level.

## Measurement of Video Modulation

Video is transmitted by being amplitude modulated onto an RF carrier. After modulation, one sideband is partially eliminated in order to conserve spectrum. The modulation format is such that sync tips correspond to the maximum carrier. Because of this arrangement, we refer to the depth of video modulation, defined as the percentage of the total amplitude change of the carrier, as the signal progresses from sync tip to white. Standard video modulation requires that 87½% = 12½% of the maximum.

The simplest (and least accurate) method of determining video depth of modulation, is to display the modulator output on a TV set, and adjust the modulation control until the picture looks good. Comparision with an off-the-air signal helps provide a reference. Unfortunately, this method is very subjective and will not yield very accurate results. It should only be used in an emergency, when no other method is available.

A second method of measuring depth of modulation is to connect a high frequency oscilloscope to the modulated IF signal and view the RF envelope directly. Under certain conditions, this yields a reasonably accurate measurement, but several circumstances can lead to wrong conclusions. For example, if the spectrum of the IF signal being examined has been passed through a vestigial sideband filter to eliminate part of one sideband, then high frequency information (primarily color information) will not be shown at the correct amplitude. This can be overcome by examining the signal before the vestigial sideband filter at the modulator, or after the Nyquist slope filter in the demodulator. Unfortunately, if harmonics of the IF signal are present at the point of measurement, these can render the oscilloscope display meaningless. Another drawback to the IF monitor approach is that the display presented is relatively difficult to interpret, especially for a color signal. Also, this requires a rather expensive oscilloscope to obtain the required 50 MHz flat response.

The method of measuring video depth of modulation most often found in CATV practice, is the use of modulation meter. This is a meter that measures the peak-to-peak amplitude of a video signal, and expresses the result as depth of modulation. The video signal may be obtained prior to being applied to the modulation circuit, or it may be obtained by demodulating a sample of the modulator output. Although more expensive, the latter approach is preferred because accuracy will not be affected by variations in modulation sensitivity. Also, by demodulating the signal for metering purposes, a failure in the modulation circuit would be readily seen.

A modulation meter is reasonably good for routine checks, but it has several drawbacks that limit its usefulness. For example, the frequency response of the metering circuit may not be flat over the entire range of video spectrum. The meter depends upon the operation of a peak detector which must accurately detect very short peaks, and must hold that peak over at least one field of the picture. A fairly complex (and expensive) circuit is required to do this. Another drawback of using a meter to measure video modulation is that some types of distortion which may accompany the modulation process, are masked. These include sync compression and incorrect operation of the peak white clipper.

The most satisfactory method of measuring video modulation depth is through use of a demodulator having a zero chopper. The zero chopper periodically switches off the picture carrier to simulate 100% modulation. This level and the sync tip level, representing 0% modulation, provide

the scaling required to measure depth of modulation. Figure 1 shows the output of a demodulator with a zero chop pulse generated during the vertical interval. A waveform monitor with a modulation depth graticule is used to display the output. By setting the sync tips on the 0% line and the chop pulse on the 100% line, the actual depth of modulation can be read directly from the scale. If such a graticule is not available, a regular oscilloscope may be used. If the sync pulse and the zero chop pulse are set eight units apart, then each unit will represent a modulation depth of 12%. Normal modulation depth will then be represented by seven divisions.

By measuring depth of modulation with a demodulator, other modulation faults, such as sync compression and improper operation of the peak white clipper, are apparent. Of course, the demodulator is also available to make many other measurements of modulator performance, as well as of off air signals.

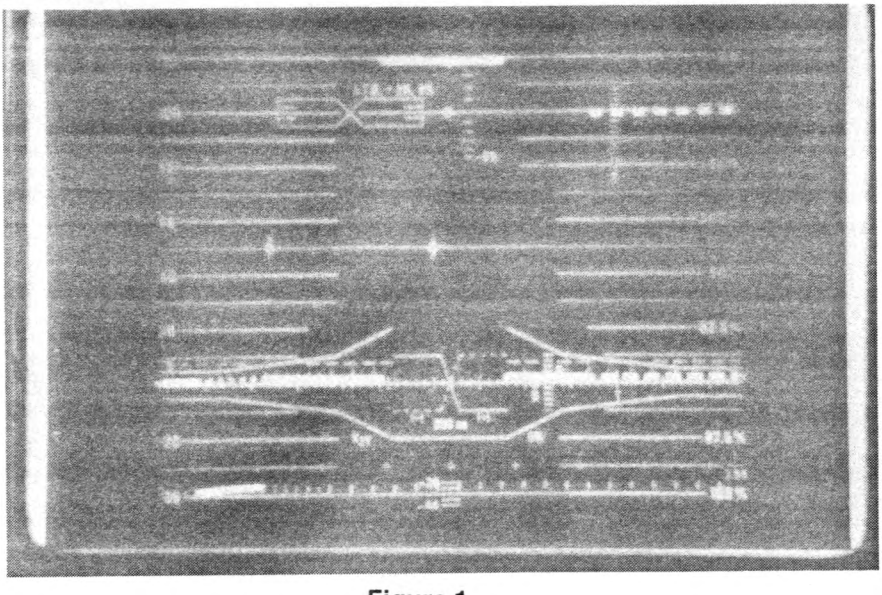

**Figure 1.**

To avoid the necessity of purchasing input converters for each channel to be monitored, the demodulator may be configured to accept an IF signal. The modulator IF output, taken just before the output converter, may be used for demodulator input.

Unfortunately, the zero chopper may itself exhibit errors in measuring depth of modulation. These errors are related to the non ideal transfer function of the detector diode in an envelope detector. Designers have used several techniques to overcome this problem, including operating the diode at very high signal level, switching a calibrating signal to the video amplifier during a chop, and biasing the detector diode. Detector problems may be overcome completely by using a synchronous detector. Here the injected carrier is used to overcome problems with the diode response. If an envelope detector demodulator is used for measuring modulation, then its response should be first calibrated using an alternate measurement.

While we are on the subject of measuring depth of modulation, one more question merits attention. This is the question of whether the normal 87½% modulation depth should or should not include the color subcarrier. Often this is a moot point because most objects of high luminance have relatively little color saturation, resulting in little color subcarrier at the white level.

However, exceptions do exist and must be dealt with. The author is familiar with one such exception. We call it the Big Bird Syndrome, named after the bright yellow character of Sesame Street fame. A case developed where trouble was observed in a broadcast transmitter, but only when Big Bird was on the set. The problem was finally traced to a video processing amplifier which clipped the peaks of Big Bird's chroma signal, generating second harmonics that were not filtered by the vestigal sideband filter. These second harmonics caused adjacent channel interference.

Broadcast engineers don't seem to be in agreement about whether or not to limit chroma peaks to 87½% modulation depth. The author's feeling is that modulation should be limited to 87½% including chroma. This minimizes TV set differential gain and phase problems, and avoids having chroma amplitude reduced, with a corresponding drop in luminance level, by the modulator peak white clipper.

**Maintenance of Video Modulation**

Having discussed the measurement of modulation depth, we turn now to the problem of maintaining that modulation. If the amplitude of the incoming video signal remains fixed, and if the modulator is stable, then no problem exists. However, video levels may vary from time to time. The broadcast industry has developed several techniques for dealing with this problem, which find application under different circumstances. If the problem is one of maintaining constant output from one or more cameras in the studio, then it is desired to monitor both the black and white signal levels, forcing them to be consistent. This may be done with a video clamp to maintain a constant black level, and means to vary the white level. White level adjustment might be accomplished at the camera by varying the iris, or might be done electronically by varying video gain. Of course, the above signal must not include sync, and the sync must be added in after adjustment. This technique is best left as a studio tool, because it is inappropriate for some programs. For example, if automatic level correction was used on a night scene, the circuitry would attempt to make it a day scene.

If it can be assumed that at some point the ratio of video to sync has been properly established, and that this ratio has not been altered by subsequent video processing, then another technique can be used. The sync pulse amplitude is examined, and the amplitude of the composite video signal is adjusted to maintain proper sync amplitude. This will assure that the video modulation depth remains as intended, regardless of the maximum or minimum luminance level in a particular scene. This technique is sometimes used at a modulator following a microwave link used to import a distant signal. Should the video amplitude provided at the output of the microwave receiver vary, consistent modulation depth could be restored.

A third method of control being used at some broadcast installations, involves tagging the problem with a reference signal at the point of origination. This vertical interval reference signal (VIRS) is theoretically transmitted through the entire transmission path, receiving the same alterations as the video signal. It may then be used to control several parameters of the transmitting equipment, including depth of modulation. While theoretically an excellent (though relatively costly) technique, the author has been told that it doesn't yet live up to its promise because the video amplitude is not always established correctly at the point of origin, and because some points in the transmission path may inadvertently strip off incoming VIRS and retransmit a second VIRS not necessarily related to the first.

## Measurement of Audio Deviation

Let us now turn from the problems of measuring and controlling video modulation to the like problems in the audio channel. Again, the problem is one of holding as close as possible to an established standard, in our case the FM deviation of 25 kHz, which should not be exceeded at any frequency.

If this deviation is exceeded, the signal will sound excessively loud and distorted. In addition, excess deviation can cause the sound carrier to rise out of the video path sound trap in the receiver. This in turn can cause the generation of 920 kHz beats in the picture, which will appear with every excess modulation peak. Under deviation of the audio subcarrier will result in weak audio and a poorer audio signal to noise ratio. Several requirements are placed on the meter used to measure audio deviation. Obviously, the first requirement is that the meter must have good static accuracy: i.e., when a single tone is applied to the modulator input, and the modulator is adjusted for an indication of 25 kHz deviation, then the carrier must actually be deviated 25 kHz. This can be established by measuring the sound subcarrier with a calibrated deviation meter. An alternate technique makes use of a spectrum analyzer. The modulator is supplied with a tone of known frequency, and the deviation control adjusted until the spectrum analyzer indicates that the carrier amplitude has dropped to zero. The lowest deviation at which this happens is that for which the modulation index (ratio of peak deviation to modulating frequency) is 2.4. [1]Since the modulating frequency is known, the deviation may be calculated.

As in the case of video metering, the signal for deviation metering may be taken either before the modulation process, or may be obtained by demodulating the audio subcarrier. The latter technique would be preferred, as variations in modulation sensitivity will not affect accuracy. In either case, though, the signal used for metering the audio should rceive the same pre-emphasis that the signal supplied to the modulated stage receives. Failure to do so will result in loss of accuracy if the highest amplitude signal component has a frequency greater than about 2 kHz. (With the 75 $\mu$s pre-emphasis time constant used in North America, the signal gain is raised by 3 dB at 2.122 kHz, and is raised by about 17 dB at 15 kHz.)

To this point, we have concerned ourselves only with the static properties of the audio deviation measurement. However, the dynamic properties of the meter are quite important. Considerable effort has been expended toward the goal of measuring audio signal level in a manner such that all audio sources that measure the same, will sound equally loud to the listener. The conventional V.U. meter widely used today, is one such attempt. However, it falls short of this goal. [2]Also, in measuring the deviation of the audio modulator, our goal is not just to meter sound for consistent loudness, but to achieve the maximum possible deviation without exceeding 25 kHz. This requires a meter whose dynamics are such that the peak level will be displayed.

[1]Therman, F.E., *Electronic and Radio Engineering,* 4th edition, 1955, p. 589.

[2]Tremaine, Howard M., *Audio Cyclopedia,* 2nd edition, 1969, p. 450.

Since this peak level will exist for only a short time, an impractically fast response time is required from the meter. This means that the meter must be driven from an electronic circuit that compensates for the necessarily slow response of the meter movement. The required circuit rectifies the audio waveform and holds the peak value long enough to permit the meter movement to rise to indicate the correct value. This gives rise to a meter movement whose characteristics differ considerably from those of a conventional VU meter. Upon application of modulation, the meter begins moving toward the appropriate indication. When the modulation amplitude begins dropping, the meter will follow, but with a sluggish response. A peak reading meter has been adopted as standard in the European Broadcast Union. Schmid in the United States has compared its performance to that of a VU meter. [3]He has found it to permit much more accurate modulation monitoring.

One additional requirement placed on the dynamics of the audio modulation meter, is that the overshoot exhibited be low. That is, when the meter reaches the final reading, it should stop quickly. All meter movements will exhibit some overshoot, but the amount should be minimized by proper meter selection and by proper matching the electronic driver.

The above dynamic properties may be explored on a given modulator by supplying the output of an audio signal generator (usually set to 400 Hz) to the audio input on the modulator, with a telegraph key in series. The key may be one of a number of hand keys used by amateur ratio operators for code transmission. When the audio is chopped into a series of short bursts ("dits"), the meter should read the same as when the key is held down. If the key is suddenly depressed and held for a few seconds, the overshoot exhibited by the meter may be studied.

A somewhat more complex test was performed on several modulators sold to the CATV industry, and also on a distortion analyzer whose meter dynamics agreed with the standard VU characteristic. All modulator inputs were connected in parallel. An oscilloscope and the distortion analyzer were supplied audio through a pre-emphasis network. Tapes used were recorded from commercial radio broadcasts in Atlanta, and were spliced into endless loops so that the same segment could be studied on each indicator. First a 400 Hz tone from an audio oscillator was supplied to all indicators, and levels were adjusted so that all modulators indicated 25 kHz deviation, the distortion analyzer indicated 0 dBm, and a reference trace was established on the oscilloscope.

The output of the tape recorder was then substituted for the oscillator. Only the recorder output level was adjusted in the tests below, so all indicators received the same peak signal. For every test tape, the recorder output was adjusted for the same peak output, as indicated on the oscilloscope. On subsequent passes of the tape, each modulator meter and the distortion analyzer (with VU meter characteristics), were checked to see what deviations were indicated. The results are tabulated in Table 1. Errors in deviation have been translated to decibels.

[3]Schmid, Hans, *"Audio Program Level, the VU Meter, and the Peak-Program Meter." IEEE Transactions on Broadcasting,* Vol. BC-23, No. 1, March 1977, p. 22ff.

**TABLE 1. METER RESPONSE TO VARIOUS TESTS**

| Test | Modulator A | Modulator B | Modulator C | Modulator D | VU Meter |
|---|---|---|---|---|---|
| 1 | −2.8 | −7.5 | 0 | 0 | −7 |
| 2 | −1.9 | −5.7 | +0.7 | 0 | −6 |
| 3 | −1.9 | −5.7 | 0 | −0.2 | −4 |
| 4 | −0.7 | −7.0 | 0 | 0 | −5 |
| 5 | −1.5 | −5.0 | +1.3 | 0 | −4 |
| 6 | +1.9 | −18.0 | −6.4 | −1.1 | |

The material on each tape is briefly described as follows:

Test 1—Male voice PSA, easy listening style.

Test 2—Heavy, explosive sound effects (from movie advertisement).

Test 3—Strong male voice.

Test 4—Male newscaster.

Test 5—Piano solo in higher octaves - fast notes with sustain pedal.

Test 6—Not really a tape, but included for comparison. This is a test of meter frequency response at 15 kHz. The signal level was reduced to take into account the pre-emphasis.

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Note first from Table 1, the inconsistency in peak reading provided by the VU characteristic. This inconsistency would probably have been even greater had not all test tapes been recorded from processed audio. Often it is assumed that a VU characteristic meter will read about 8 dB below peak level on voice, with a somewhat lower error on music. Note next that modulators A and B both read low on all tests. In neither case was the meter movement driven by a peak detecting circuit. The meter movement in Modulator A was sluggish but exhibited overshoot, which tended to make the readings look a little closer to what they should. However, the overshoot can cause misleading indications in the opposite direction. In a preliminary test, this modulator indicated too high at times when the audio input was a female voice on a soap opera. The meter on Modulator B exhibited less overshoot, but it too was sluggish and failed to indicate true peaks. In addition, the audio signal supplied to the meter circuit did not include pre-emphasis, as indicated by Test 6. Either of these meters would be satisfactory if used only to set levels using a reference tone, with program level set elsewhere.

However, only Modulators C and D contained meters that would permit the correct deviation to be set using normal program material.

## Maintenance of Audio Deviation

We come now to the final topic to be discussed in this paper: techniques used to maintain the maximum permissible deviation without exceeding 25 kHz. If the object is to maintain the highest possible deviation, then automatic modulation adjusting equipment must take into account the pre-emphasis of the modulator. When we see audio processing equipment designated for FM, this is usually the reason: The equipment first pre-emphasizes the signal, processes it, then de-emphasizes. Thus, the signal is processed with the same pre-emphasis that it will receive at the modulator.

Another precaution which must be observed when using audio processing equipment that is separate from the modulator, is the introduction of group delay between the processor and modulator. Group delay can shift the relative phase of different frequency components of a complex waveform, so that the peak amplitude of the signal could be changed after processing. This is of concern primarily to the sterophonic F.M. broadcaster who employs a 19 kHz trap in the input to the stereo exciter. However, it is mentioned here as a precaution should a selective filter ever be necessary between an audio processor and a modulator.

Having taken the above detour, we return to the question of audio processing to maintain maximum legal deviation. Audio processing generally falls into one of three classifications: compression, modulation limiting, and clipping. Within each of the first two categories, we can talk about several variations. These techniques and variations appear to comprise the various audio processing approaches available. We'll deal here with each, discussing suitable applications.

Compression is the term generally applied to the reduction of the entire dynamic range of the program material, i.e., "riding the gain" automatically. The object is to maintain either a reduced dynamic range, or no dynamic range, in the output level. This technique is quite proper and popular for processing speech, where it can reduce such problems as level changes when a speaker turns away from a microphone, or when a new speaker begins talking. Compression is also used with rock music and other music formats in which dynamics are not a part of the art. However, use extreme caution in applying compression with classical music, in which dynamics are very important. It is something to behold the wrath of a classical musician whose work has had dynamics eliminated by over-zealous application of compression! Compression is generally specified by a ratio of N:1, meaning that N dB of input level change will result in 1 dB of output level change. A compression ratio of 2:1 would represent fairly small "meddling" with the dynamic range, while 10:1 would be a fair amount of compression.

Other criteria apropos to compression include attack and release times, and a related decision of whether the output peak amplitude shall be monitored, or whether output rms or average level (taken over some time span), is to be maintained. A variation available appears to involve equipment that separates the audio spectrum into two or more bands, compressing them individually.

Modulation limiting is the second technique for audio processing. Unlike a compressor, a modulation limiter does nothing to the audio level until a threshold (25 kHz peak in TV audio) is reached. Above this threshold, the limiter acts as a compressor with a very high (greater than 20:1) compression ratio. Thus, dynamic range is unaffected until a peak attempts to overdeviate

the carrier. At this time, the gain is reduced until the modulation returns to a lower level. Modulation limiting is not to be confused with RF limiting, which clips the RF peaks, generating distortion. Modulation limiting is simply an automatic turning down of the gain if the level gets too high. This technique is often used as protection at a transmitter, to prevent overmodulation without introducing distortion. It is safe to use with all program material if applied properly.

If the modulation limiter is overdriven, it acts as a compressor with very high compression ratio. This may be desirable in some instances, where a limiter is intentionally overdriven by a modest amount (8 dB seems to be a common figure) during normal programming. This permits minor drops in level, and increases in level, to be compensated, without masking the normal dynamic range of the material.

When using either a compressor or limiter, levels cannot set arbitrarily: There always exists a maximum level that can be accommodated by circuits prior to the gain adjustment stage. The difference in level between normal operation and the maximum signal that can be handled without distortion, is called the headroom. Headroom is specified for a constant sinusoidal tone, not for complex program material. If the rise in input exceeds the headroom for the particular equipment, then distortion will occur even if the output level does remain constant.

The final type of audio processing to be discussed is clipping, a process which does cut off the extreme levels of signal voltage. Clipping introduces distortion, and thus would not be used during normal programming. A clipper is normally set to act at some deviation greater than normal (e.g., 40 kHz for TV sound). It acts only as a "court of last resort," to prevent gross overdeviation. If the signal were allowed to deviate too far, then the sound carrier would begin riding out of the sound trap preceding the video detector in the receiver, giving rise to 920 kHz beats in the picture.

Summarizing the three types of audio processing, a compressor is useful for a range of audio problems, but if misapplied can be detrimental to the overall effort. Because the right amount of compression is a function of the type of programming, this is best left as a studio technique. The engineer concerned with signal modulation and transmission, should concentrate while not appreciably affecting normal program audio. As a final clean-up measure, a clipper may be used to prevent accidental gross overmodulation, but it should not affect normal program audio. When using a modulation limiter, a clipper is not as important as when no protection is supplied. However, since modulation limiters have finite attack times, then a clipper could occasionally be needed to clip one or a few cycles of the audio, before the limiter acts.

## Conclusion

We have tried to mention some of the considerations apropos to operating television modulators. While financial and operating situations may exclude some of the techniques discussed, the ideas presented should be considered when problems arise or equipment upgrading is contemplated. Fortunately, no amount of electronic wizardry will replace a sharp technical person, who knows what techniques to apply where.

# SECTION 5
# 416 TELEVISION MODULATOR

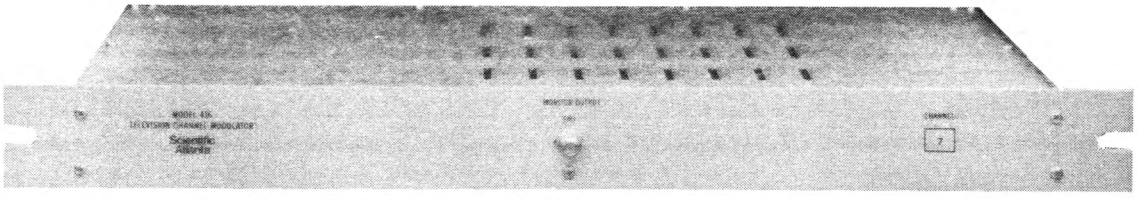


**Figure 5-1. Model 416 Television Modulator**

## Description

The Series 416 is a basic Television Modulator that delivers Scientific-Atlanta's traditionally high quality at an economical price. This product was designed for use where the need for flexibility is minimal, but the demand for quality and reliability in CATV, MATV, and ETV systems is critical.

The Series 416 Television Modulator accepts an audio and video signal and generates a high quality vestigial sideband television signal. VHF, midband, and superband channels are available.

Surface acoustic wave (SAW) technology is used to provide outstanding vestigial sideband response characteristics with superior adjacent channel performance and excellent group-delay characteristics. Vestigial sideband filter adjustments are virtually eliminated by the SAW filter, and temperature stability is significantly improved over conventional methods.

In order to conform to standard FCC pre-distortion requirements, a delay pred stortion network is provided as standard.

A peak white-level clipper limits video modulation to 95%, eliminating the possibility of carrier cutoff or phase reversal, a major cause of sync buzz.

Automatic phase-lock circuitry provides accurate video-to-sound carrier separation.

The standard version Series 416 is powered by, and operates as an accessory to the Series 6600 Video Receiver. An optional AC powered version enables the modulator to operate with any receiver. Both have a front-panel RF output port for monitoring or testing.

## Options

- IF Loop Thru (IFL)
- AC Powering (AC)

# Model 416 Television Modulator Specifications

**VIDEO INPUT**

Input
- Type: Negative sy
- Level: 1V p-p for 87.5% modulation depth
- Impedance: 75 ohms, unbalanced
- Return Loss: 26 dB

Frequency Response
- ±1 dB, 15 Hz — 4.2 MHz

Differential Gain
- ⩽5%, max modulation at 87.5%

Differential Phase
- 1°, max modulation at 87.5%

Hum and Noise
- −60 dB with respect to 90% modulation

**AUDIO INPUT**

Input
- Level: 0 dBm for 25-kHz peak deviation at 1 kHz
- Impedance: 600 ohms, balanced
- Return Loss: 26 dB

Frequency Response
- ±1 dB, 30 Hz — 15 kHz

Pre-Emphasis
- 75 μs

Harmonic Distortion
- 1% max with 25-kHz deviation at 1 kHz

FM Hum and Noise
- ⩽60 dB with 25-kHz deviation at 1 kHz

**RF OUTPUT**

Channels Available
- 2-13, A-W

Output
- Level: +50 to +60 dBmV, continuously variable
- Impedance: 75 ohms, unbalanced
- Return Loss: 16 dB
- Output Level Stability: ±1 dB

Intercarrier Level
- Audio carrier 15 dB below video carrier (adjustable)

Intercarrier Frequency Tolerance
- Audio carrier 4.5 MHz ±500 Hz above video-carrier
- ±10 kHz

Vestigial Sideband Response
- At Channel Edge: −20 dB
- At Adjacent Picture and Sound Carriers: −40 dB

Spurious Output
- ⩾60 dB below video carrier (40 — 300 MHz)

**RF Output Monitor**

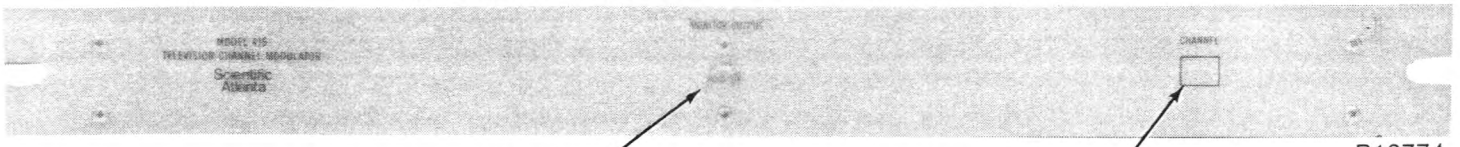
Output
- Level: −30 dB
- Impedance: 75 ohms, unbalanced
- Return Loss: 10 dB

**GENERAL**

Power Requirements
- Standard: +15V dc, 400 mA, −15V dc, 400 mA
- Optional: 105 — 130V, 47 — 63 Hz, 25W

Dimensions
- Standard 483-mm (19 in.) EIA panel, 44 mm (1.75 in.) high, 343 mm (13.5 in.) deep

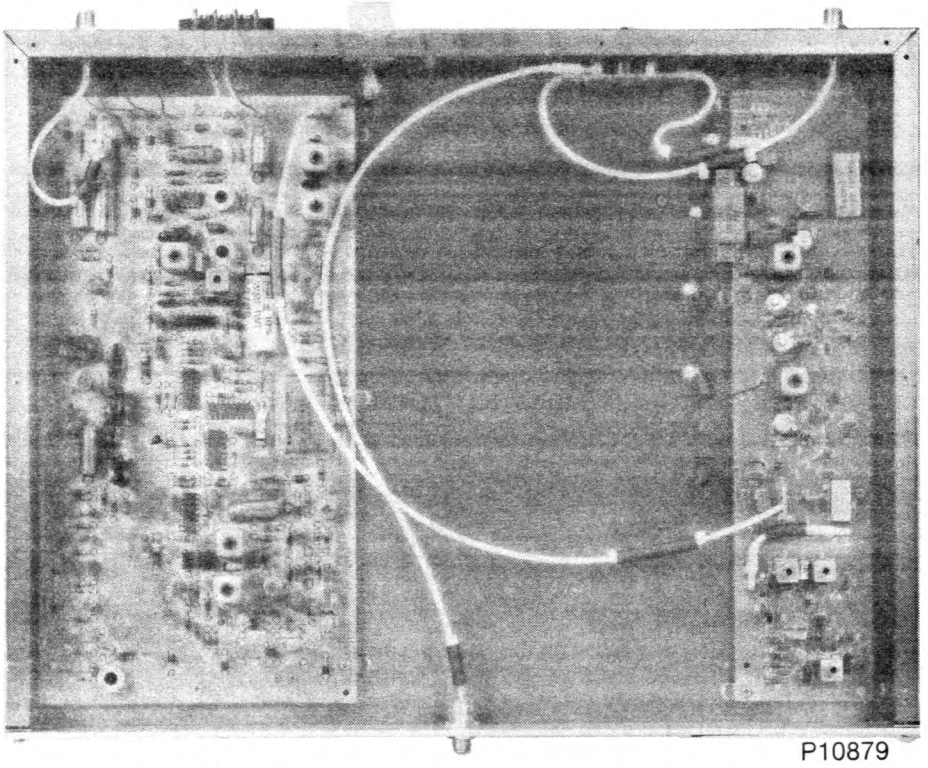
Weight
- Net: 4.5 kg (10 lb)
- Shipping: 5.5 kg (12 lb)




P10774

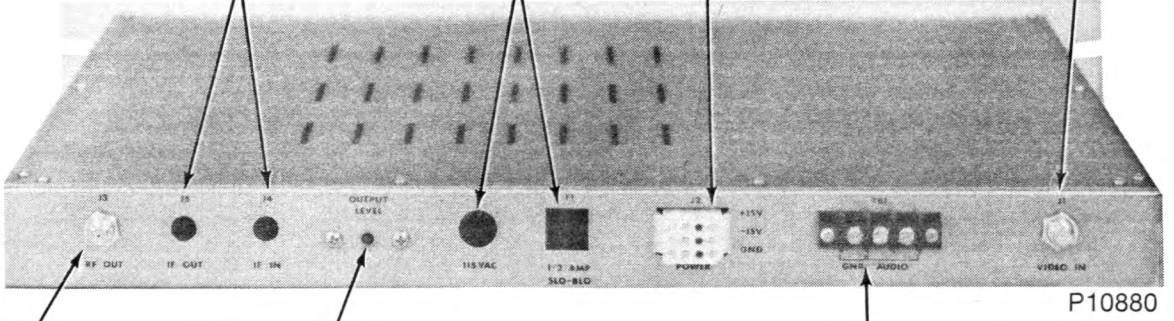
−30 dB
RF OUTPUT
COUPLER

TELEVISION CHANNEL
AT RF OUTPUT



P10879

**Figure 5-2. Television Channel Modulator, Top View**




P10880

OPTIONAL

USED ONLY
ON 416A

DC PWR
INPUT

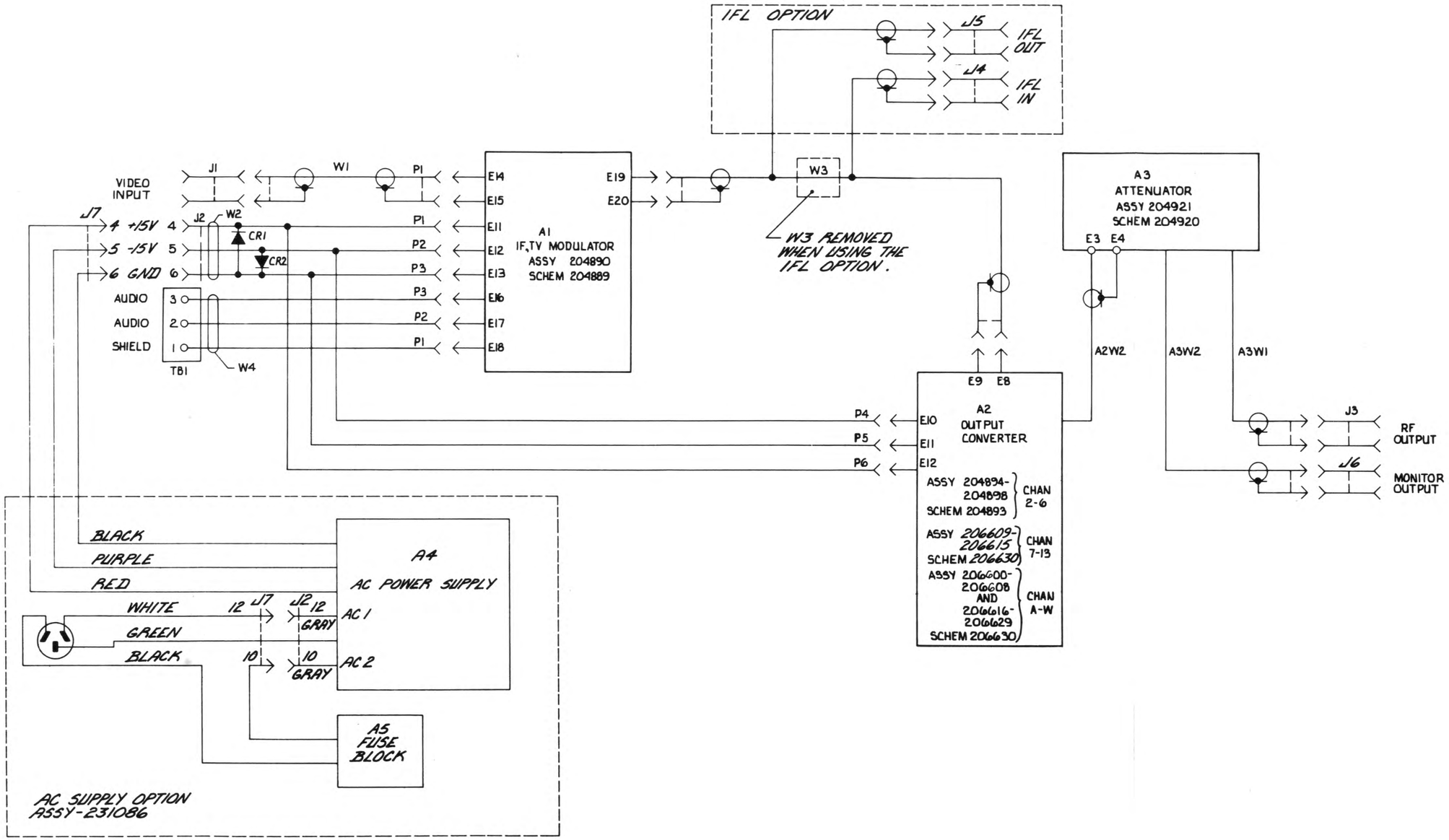
TYPE F
75 OHMS UNBALANCED
1V p-p INPUT
FOR 87.5%
MODULATION

75 OHMS
UNBALANCED
TYPE F
MODULATED
RF OUTPUT

THE OUTPUT LEVEL
ADJUSTMENT WILL
ADJUST THE RF
OUTPUT FROM
+60 dBmV TO +50 dBmV

600 OHM BALANCED
0 dBm AT 1 kHz FOR
25 kHz PEAK DEVIATION
AUDIO INPUT

**Figure 5-3. Television Channel Modulator, Rear View**




D204882D

**Figure 5-4. Series 416 Television Channel Modulator Schematic Diagram**

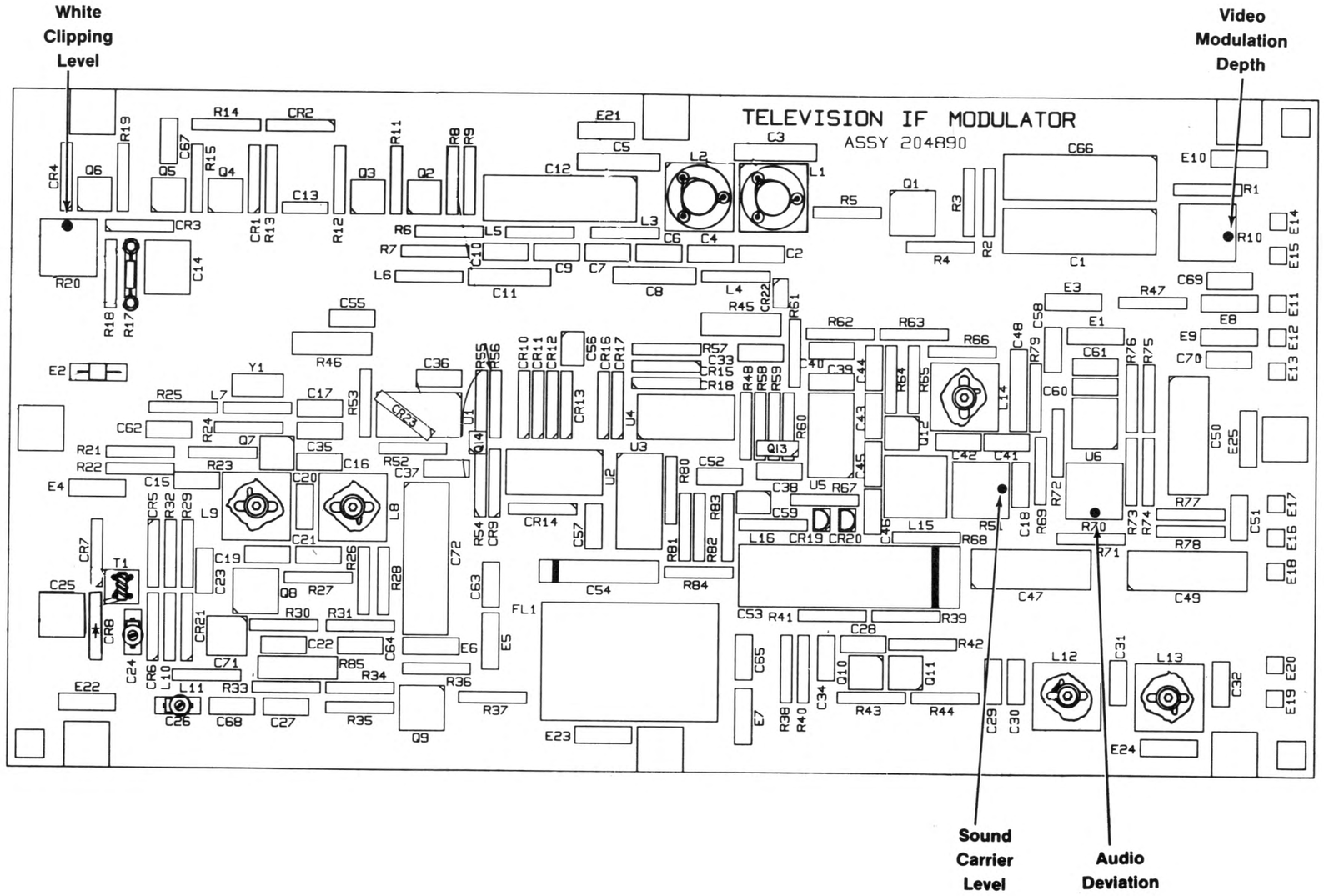
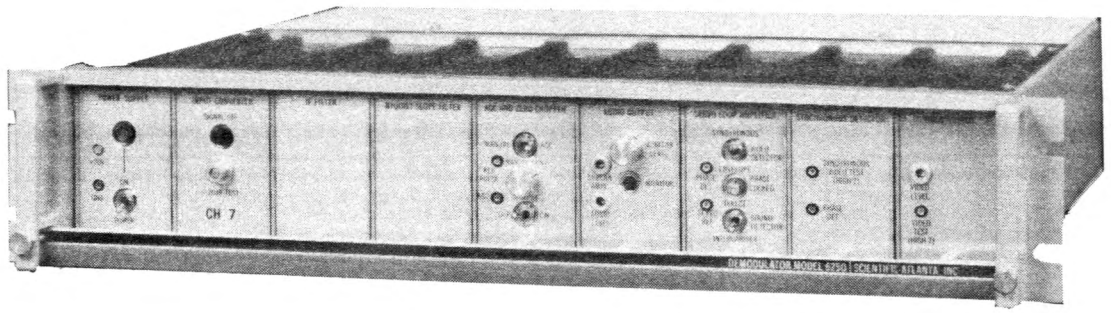



**Figure 5-5. 416 Modulator Internal Adjustments**

# SECTION 6
# 6250 TELEVISION DEMODULATOR




Shown with optional synchronous detector.

**Figure 6-1. Model 6250 Television Demodulation**

## Options

- Synchronous video detector is free of quadrature distortion that is found in all vestigial sideband envelope detectors. With the synchronous detector, capability is provided on the front panel to rapidly switch between envelope and synchronous detectors, greatly enhancing the diagnostic and operational capabilities of the demodulator. If the synchronous detector is supplied, two options are available in the sound detector. The source for the 4.5 MHz aural subcarrier can be switched between either a direct mixer, using the phase-lock synchronous oscillator and audio IF signal, or an intercarrier detector using composite video and sound IF signals. (SK — add on kit)
- 50 ohm input impedance (MT)
- Carrier Loss Relay (SR)
- Low Level IF Switching (LIFS)
- Automatic IF Sensing (AIFS)
- Audio IF Override (AO)
- Standby Battery Powering provides backup for failure of 115V ac. (SP)
- Negative dc powering permits operation from a −19 to −31V dc battery supply in lieu of 115V ac. (DC)

## Description

The Model 6250 Television Demodulator provides precision demodulation of any transmitted television signal. The envelope delay provided closely matches that of the average home television receiver. This permits use with transmitters which have been predistorted to compensate for this delay. An integral zero chopper enables measuring the depth of modulation of the transmitted signal. In addition, a programming option allows external command of the zero chopper.

The demodulator is equipped with an envelope detector (ED) for the video, and may be ordered with an optional synchronous detector (SD). The synchronous video detector is free from quadrature distortion that is found in all vestigial sideband envelope detectors. Quadrature distortion is a source of unsymmetrical ringing and smear and cross talk between video sideband components.

Audio is provided both as a baseband signal and as a 4.5 MHz subcarrier. A loop-through is provided to permit the addition of the aural subcarrier to one of the video outputs for transmission of the sound signal at a frequency above the spectrum occupied by the video signal in microwave installations.

# Model 6250 Television Demodulator Specifications

Except as noted, specifications apply to both synchronous and envelope detectors and to both direct and intercarrier sound detection. These specifications apply to demodulators using single channel input converters, and may not apply when tunable input converters are used. Input levels refer to 75 ohm input impedance. Multiply by 0.8 for 50 ohm input impedance.

Overall Sensitivity
.1 mV (minimum input for 1v p-p video output)
Input Frequency
Any standard VHF, UHF, sub-low, mid-band, super-band, IF, or non-standard offset frequency
Input Level Dynamic Range
.1mV to 32.8mV (−20 dBmV to +30 dBmV), VHF
.1mV to 17.5mV (−20 dBmV to +25 dBmV), UHF
Input Impedance and VSWR
50 or 75 ohms (specify), VSWR less than 1.35:1 over channel of interest, over entire dynamic range of input
Noise Figure (at maximum gain)
6 dB Low Band
7 dB High Band
9 dB UHF
Image Rejection
60 dB VHF
50 dB UHF
IF Rejection
60 dB
IF Frequencies
Video — 45.75 MHz and Audio — 41.25 MHz
IF Frequency Response
Follows standard Nyquist response
Adjacent Channel Rejection
60 dB
Video Amplitude Frequency Response
10 Hz — 4.18 MHz ±.5 dB
Envelope Delay
Complement of standard transmitter group delay pre-correction
10 Hz — 3 MHz ±40 ns
3.58 MHz + 170 ±20 ns
Differential Gain
±2% (synchronous detector)
±2.5% (envelope detector)
Differential Phase
±0.5° (synchronous detector)
±1° (envelope detector)
Chronimance-to-Luminance Crosstalk
3% (envelope detector)
Unmeasureable (synchronous detector)
Above values are percentage of blanking-to-white-level luminance shift as chrominance signal varies from 0 to 100 IRE units, 87.5% modulation depth
Zero Chop Position, Length
Start time adjustable from 3 - 6 lines after vertical sync pulse, width adjustable 1 - 3 lines
Video Output Level
Variable 0 - 1.5V p-p at each of two outputs
Video Output Impedance
75 ohms (30 dB min. return loss)
Squelch
On — delayed approximately 10 seconds
Off — instantaneous; adjustable threshold
Sound Output Levels
Audio
+11.5 dBm max across 600 ohms
Aural Subcarrier
0.2V p-p max across 37.5 ohms
Above levels adjustable down to 0
Audio Frequency Response
±0.5 dB, 30 Hz - 15 kHz with de-emphasis
De-Emphasis Time Constant
75 μs, ±2%
Audio Harmonic Distortion
1% max at any frequency, maximum deviation and output level
Audio Monitor
Standard phone jack with level control — will drive most headphones

**GENERAL**

Input Voltage
100 to 130V ac, 50 to 60 Hz
+24V dc nominal standby battery with optional built-in automatic switchover
−24V dc nominal battery operation with optional dc converter module
Input Power
30 watts maximum
Weight
10.4 kg (23 pounds)
Mounting
89mm (3½")H x 483mm (19")W x 495mm (19½")D (standard rack mount spacing; rack slides optional)
Operating Temperature Range
−7 to +50°C (+20 to +120°F)

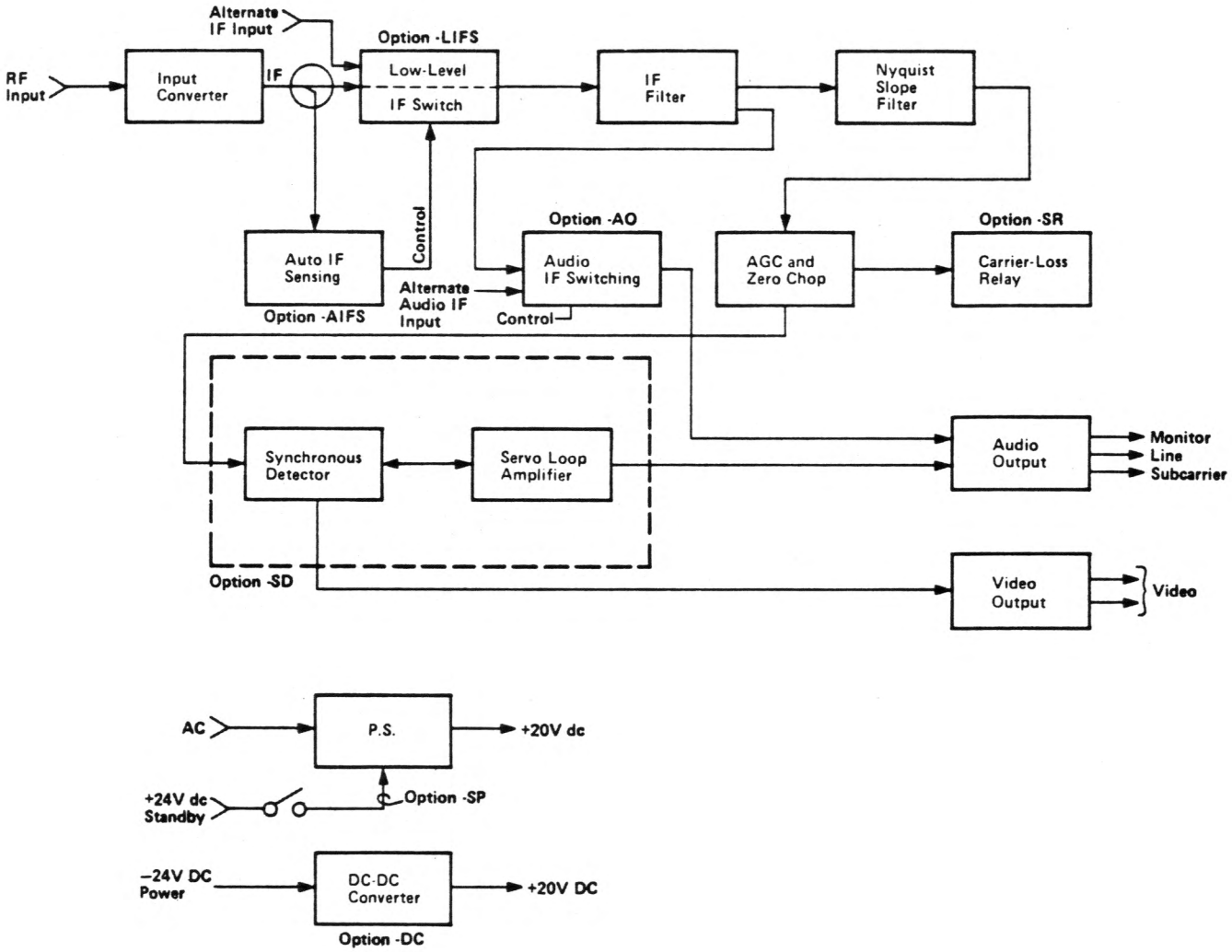



**Figure 6-2. Model 6250 Block Diagram**

# IMPROVED HEADEND PERFORMANCE USING A TEST DEMODULATOR

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Besides its traditional applications, a precision demodulator can benefit the CATV engineer as a tool for improving system performance. In the past, demodulators have been used as inputs to microwave links, and have found use in demod-remod types of headends. In addition to these applications, a demodulator can be of great value during modulator evaluation and set-up, in aligning antennas, and in evaluating headend equipment on an in-service basis.

Before entering into a discussion of these applications, a brief review of some current debate concerning demodulator characteristics is in order. The two primary questions meriting attention are the choice of envelope vs. synchronous detection, and the most desirable shape of the IF passband response. This latter question is of concern primarily as it applies to envelope detectors. Both of these questions are currently being debated within the broadcast/TV receiver/CATV industry, so a divergence of opinion may be noted.

The question of synchronous vs. envelope detection is probably headed for a solution (though no "solution" is ever permanent), with the conclusion that both types of detectors are necessary if a reasonably complete evaluation of the modulated signal is to be undertaken. For operational situations (e.g., feeding a microwave relay link), either detector yields reasonable results, with the synchronous detector being preferred. Again, the presence of both detectors would be desirable.

## Envelope Detection

Virtually all television sets in use (with the exception of a few marketed within the last two or three years) employ an envelope detector. This usually takes the form of a diode after the last stage of the IF amplifier, placed in series with the output. The IF signal is rectified by the diode, reproducing the envelope of the IF signal. Envelope detectors are characterized by simplicity, low cost, and relative immunity to transmitter incidental FM. Perhaps the most compelling reason for using an envelope detector in a test demodulator is the fact that it is so common in the home receiver. Hopefully this will give us the opportunity to estimate the performance of the detector in a home receiver.

However, envelope detectors have two very significant drawbacks, one related to errors in the envelope of a vestigial sideband signal, and the other related to the practical problem of diode nonlinearity.[1] The former problem is called quadrature distortion, and arises with any signal in which the sidebands are filtered unsymetrically about the carrier.

Among the errors caused by quadrature distortion are: ringing on transitions, differential gain (though this can also rise from diode non-linearity), and chrominance to luminance crosstalk, resulting in a darkening of picture areas having considerable color.

One of the classical analyses of the quadature distortion problem is to compute the actual vs. ideal luminance and red, green, and blue levels developed during envelope demodulation of color bars. This calculation shows considerable shift in all levels.[1] However, the amount of shift is drastically reduced if the IF passband is shaped such that the color subcarrier is reduced before detection. This is the reason that the IF response of most television sets is reduced at the color subcarrier frequency.

The second problem associated with diode detectors is due to the non-linear forward response of real diodes. This is generally manifested as reduced diode efficiency at lower signal levels, resulting in additional differential gain and phase in the output, as well as additional distortion of transient responses. These problems generally are seen near the white level, corresponding to minimum signal amplitude. The magnitude of the problem can be reduced through use of high quality diodes operated at high signal levels.

### Synchronous Detection

An alternate to envelope detection is synchronous detection. A phase-locked loop is normally used to reconstruct a CW reference signal having a definite phase relationship to the incoming picture carrier. This signal is mixed with the incoming carrier, and one resultant output is baseband video.

Everything that was said about the envelope detector could be applied in reverse to the synchronous detector. The synchronous detector, due to its cost and complexity, has seen virtually no application in consumer television sets, but has been available to the professional user for about six years. Recently several integrated circuit manufacturers have introduced detector chips that claim to use synchronous detection. However, the reference signal is derived by introducing a sample of the modulated signal into an amplitude limiter. While this technique produces acceptance performance for consumer applications, our experience to date indicates that these chips are inappropriate for precision applications, and don't yield true synchronous detection.

Besides cost, complexity and lack of consumer application, synchronous detectors suffer when the transmitter imposes incidental frequency modulation on its carrier. This is especially a problem with some older transmitters, due either to insufficient power supply filtering or to insufficient power supply regulation. When incidental FM is present, both synchronous and envelope detectors suffer, though the degradation is much greater with the synchronous detector. This is due to the inability of the synchronous detector to follow the frequency modulation present. The severity of the problem is partially determined by the servo loop response of the phase-locked loop, compared with the effective waveshape of the frequency modulating signal.

On the other hand, the synchronous detector, properly adjusted, does not respond to quadrature modulation components, so all the problems mentioned above related to quadrature distortion do not exist. Also, since the reference signal may be applied to the mixer at a very high level, and since the mixer may be designed to not operate diodes near their reverse breakdown voltage, the problem of diode non linearity is largely avoided.

### Detector Choice

We thus face a choice of which detector to use. Generally, we would prefer a synchronous detector for both off-air and test applications, due to its ability to demodulate pictures more accurately, and the lack of possible masking of modulator faults.

On the other hand, the criterion by which technical performance must ultimately be judged is the picture seen by the consumer. Since most consumer sets use envelope detection, one could argue that an envelope detector would evaluate a modulator more nearly like a TV set would evaluate it. Since tradeoffs can be made in the transmitter delay predistortion to crudely com-

pensate for envelope deficiencies, transmitters doing this might yield superior results when demodulated with an envelope detector. (The author has not witnessed this, though the possibility has been discussed.)

The broadcast industry is in the process of concluding that a measurement demodulator must have both types of detection. This philosophy should surely follow for demodulators used by CATV systems for test applications. For off-the-air applications, the availability of both types of detectors is again useful: the synchronous detector to use for highest quality demodulation during normal operation, with the envelope detector as a backup to use if incidental FM is present.

### Passband Response

Another question to be answered when specifying a demodulator, is the predetection (IF) passband shape. As discussed above, improved response from an envelope detector is obtained when the IF response is reduced at the color carrier frequency. For this reason, most consumer receivers reduce the response at the color subcarrier. Unfortunately, no standards exist concerning the passband response, and receivers differ markedly in their response.[2] At this time, the industry is debating the passband response of a test envelope demodulator. In some manner, the response will probably be reduced at the color subcarrier, with post detection restoration of the lost frequency response. Of course, since quadrature distortion doesn't affect a synchronous detection, no similar problem exists with the synchronous detector. The IF response is held flat at the color subcarrier frequency and a video amplifier with flat response is used.

### Application

When a CATV system uses modulators, a demodulator should also be available to aid in measuring modulator performance. The most fundamental measurement to be made on a continuing basis is depth of modulation. In a standard television signal, a maximum modulation depth of 87.5% is used. Modulating to greater depth may result in intercarrier buzz in the receiver, as well as causing washout of the brighter areas of the picture, and excess differential gain and phase due to diode non-linearity. Modulation to an insufficient depth will create a washed out picture, loss of color brillance, and possibly even loss of sync in the receiver. Some CATV modulators have no means for measuring depth of modulation. However, the more popular modulators do have meters for indicating approximate depth of modulation.

A meter is better than nothing, but it also has limitations. It may not detect a very brief excursion of some highlight into the overmodulated region, nor may it tell if the color carrier (only) makes excessive excursion. Also, the accuracy of a meter is only 5 percent or so.

Many modulators incorporate a peak white clipper that won't permit overmodulation. However, the clipper can only be set while looking at demodulated waveform or while looking at the modulated IF with a high frequency oscilloscope. This latter technique must use the IF signal before the vestigial sideband filter, because the filter will effectively reduce the amplitude of the color subcarrier.

In using a demodulator to measure depth of modulation, two references must be established. One reference is at the maximum amplitude of the RF carrier, corresponding to the sync tips. The

other reference should be at a known depth of modulation, preferably corresponding to cutoff (100% modulation). This reference isn't available in the modulated signal, so it must be added by the demodulator. This is done with a zero chopper, a switch that interrupts the IF (or RF) signal momentarily, without affecting the AGC voltage. This interruption, or zero chop pulse, then establishes a reference.

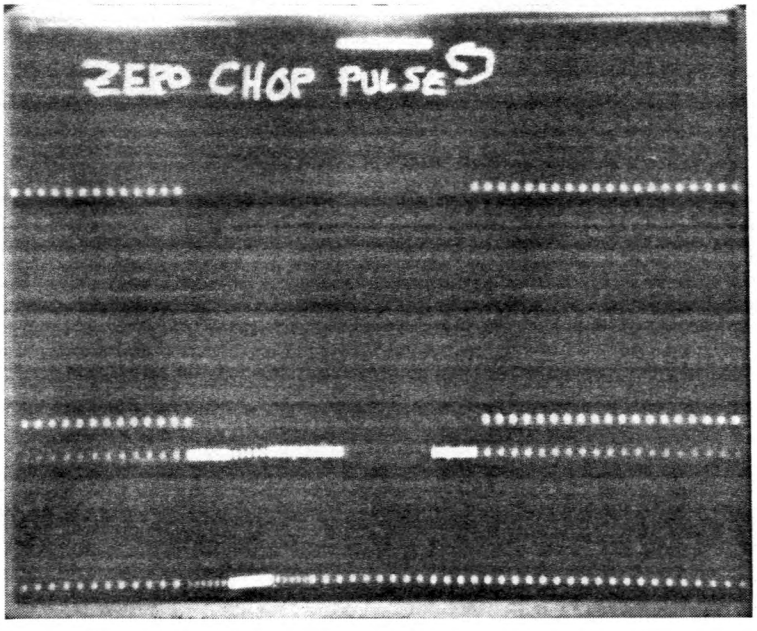


**Figure 1A**

Figure 1 shows a demodulated signal featuring a zero chop pulse occurring during the vertical interval. Figure 1A shows the vertical interval expanded, and Figure 1B shows the same signal with
the waveform monitor set for a two line display. The waveform monitor display in Figure 1A has been lowered to show the zero chop pulse more clearly. If a waveform consisting of 40 units of sync and 100 units of video is modulated onto a carrier to a depth of 87.5 percent, then the zero chop pulse corresponds to 160 units of signal. Thus, by setting the sync and chop spaced by 160 units, the sync should occupy 40 units, the video 100 units, and the zero chop pulse should be 20 units above the maximum video. A synchronous detector is definitely preferred for making this measurement, since diode non-linearity in an envelope detector will shift the zero chop level.

An incorrectly set peak white clipper can be detected as compression of the peak video (if set too low), or as permitting overmodulation if set too high.(Of course, this would best be checked during non-program times, as it would be necessary to intentionally attempt to overmodulate.)

Sync compression is another fault that could be checked using the test demodulator. If the sync and zero chop are set as above, and the blanking level is offset too low on the monitor, then sync compression may be present. Naturally, the fault could also be with the input to the demodulator, so this must be checked.

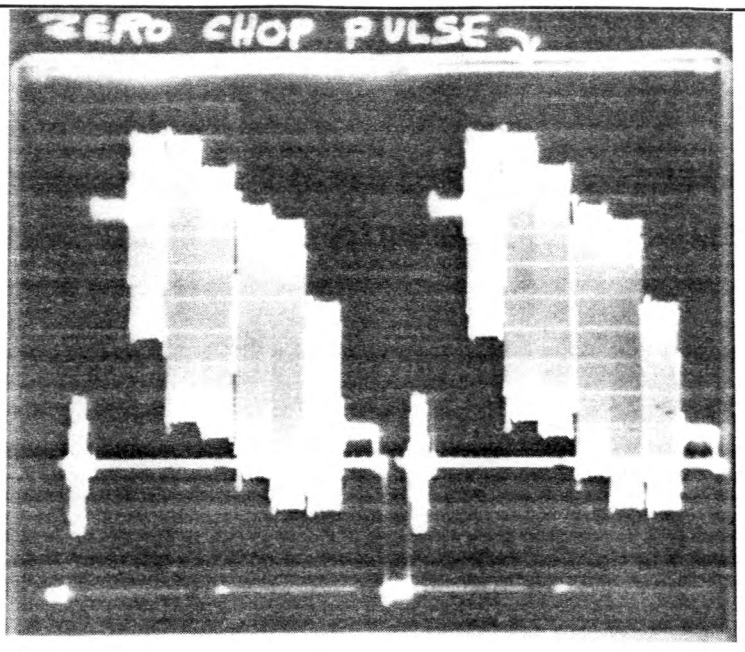


Figure 1B

Use of the demodulator also permits checking differential phase and gain in the modulator. Differential phase is a change in phase of the color subcarrier as the luminance level varies, and results in hue shifts as a function of brightness. Differential gain is a change in chroma amplitude as the luminance amplitude changes, resulting in errors in color saturation.

## Noise Measurement

A demodulator is used in conjunction with a noise measurement test set to make in-service measurements of signal-to-noise ratio (S/N). The signal is demodulated and the baseband video is supplied to noise measuring equipment. Note that the noise measured in this manner is the total noise accumulation between the point of clean video origination and the point of demodulation. Thus, the noise will include sources beyond the control of the CATV system. However, since this total noise is the noise as seen by the subscriber service. By making S/N measurements at the antenna, after the headend, and in the distribution system, a realistic picture of noise contributions can be made, permitting the weakest points in the chain to be attacked.

In making S/N measurements, it is important to distinguish between different definitions of signal and noise. In some definitions, the signal is considered to extend from the sync tips to peak video (EIA definition), while other, more common, definitions regard only the signal from blanking to white level (CCIR, BTL definitions).[3] Also, one of several weighting networks are used at times, which roll off the higher frequency noise to account for its smaller contribution to subjective picture impairment. In all cases, "signal" is defined as the peak-to-peak value (with or without sync), and "noise" is the rms value.

These definitions may all be related to the NCTA definition of carrier-to-noise ratio (C/N). For example, the relation between NCTA C/N and unweighted video S/N, where the signal is defined from blanking to white, is

$$S/N = C/N - 6.92 \text{ dB}$$

where S/N and C/N are both expressed in decibels.
An RF bandwidth of a 4.0 MHz is assumed.[4] If the signal was to include the sync, then the S/N would increase by 20 log 1/.7 = 2.93 dB relating S/N to C/N by 3.99 dB. This is the 4 dB traditionally used in CATV work.

The relation stated in equation form above should prove useful in that video S/N measuring equipment capable of in-service use, is available.[5] This equipment defines signal as being from blanking to white, and may be used with or without a weighting filter. Note that the relationship between S/N and C/N is only valid if all of the noise is attributable to sources that continue to exist after the carrier is removed. Obviously, this is not the case when noise is being transmitted from the TV station.

## Antenna Alignment

With proper interpretation, a demodulator and waveform monitor can be used to aid in alignment of antennas. In this application, only comparative readings may be used, because factors other than antenna alignment may affect results. These techniques are most effective if the station being monitored is transmitting vertical interval test signals (VITS). The photographs shown in this section were made with intentionally extreme signal distortions, generated by using a rabbit-ear antenna at different orientations.

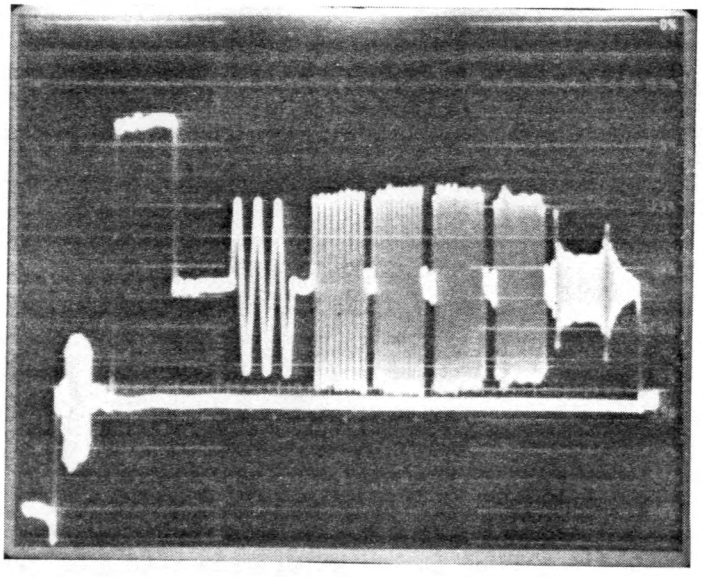

**Figure 2A**

Figure 2 shows the VIT multiburst transmitted by a well-run Atlanta VHF television station. The receiving site was about 10 miles from the transmitter. A high quality synchronous demodulator was used in making all pictures. In Figure 2A, the demodulator input was from a rooftop antenna oriented reasonably well. About 0.7 dB of peaking is observed in the middle of the multiburst. It is impossible to determine whether this is due to the transmitter, transmitting antenna, radio path, or receive antenna. Of course, the demodulator could also be at fault, except that its response was measured as flat with ±0.12 dB before these pictures were made.

The remaining photographs in Figure 2 were taken with the same test set up, except that rabbit-ears were substituted for the roof-top antenna. This was done to intentionally induce drastic changes for illustrative purposes. However, the same principles will apply during antenna alignment.

In Figure 2B, the multiburst shows a rolloff of about 5 dB from the first burst. Re-orienting the antenna yielded about 7.5 dB of high frequency peaking, as shown in Figure 2C. Even though we see some rather horrid frequency responses in these pictures, the pictures were all quite viewable to this point. In Figure 2D, the antenna is again re-oriented, with very bad peaking. This picture would probably not be considered viewable.

Figure 3 is a similarly made sequence looking at the 2T and modulated 12.5T pulses transmitted in VITS. Figure 3A is from the rooftop antenna. This would be an excellent signal, with the only fault being the undershoot at the trailing edge of the 2T pulse. Again, the demodulator was exonerated prior to the test, but this undershoot could be caused by the transmitter, either transmit or receive antenna, or the radio path. The strategy in using these tests is to orient the antenna for the best received signal. In Figure 3B, two ghosts are seen in the two small pulses following the 2T pulse. Also, the modulated 12.5T pulse shows evidence of considerable luminance to chrominance delay. Even so, the picture was quite viewable.

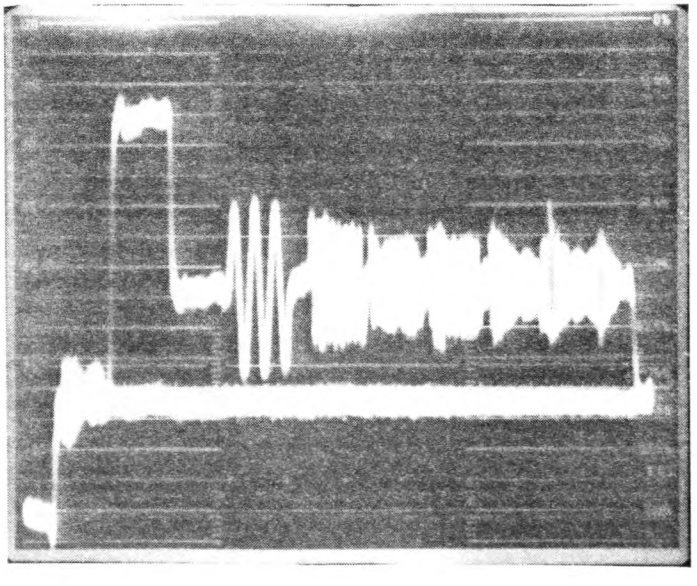

**Figure 2B**

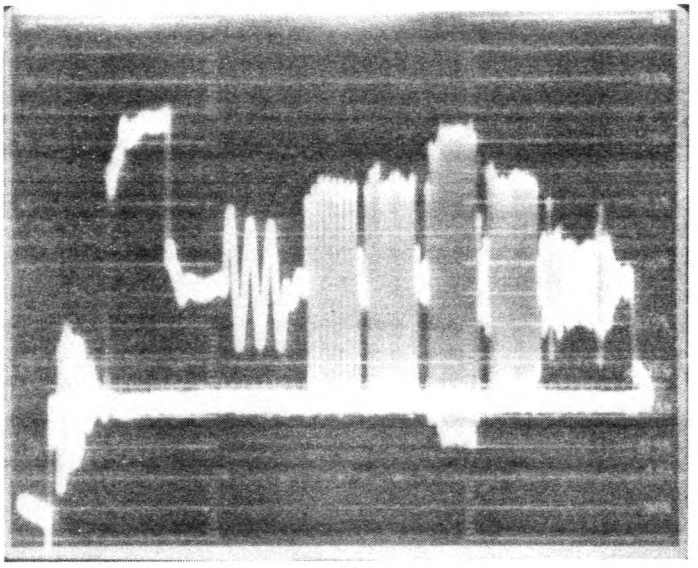

Figure 2C

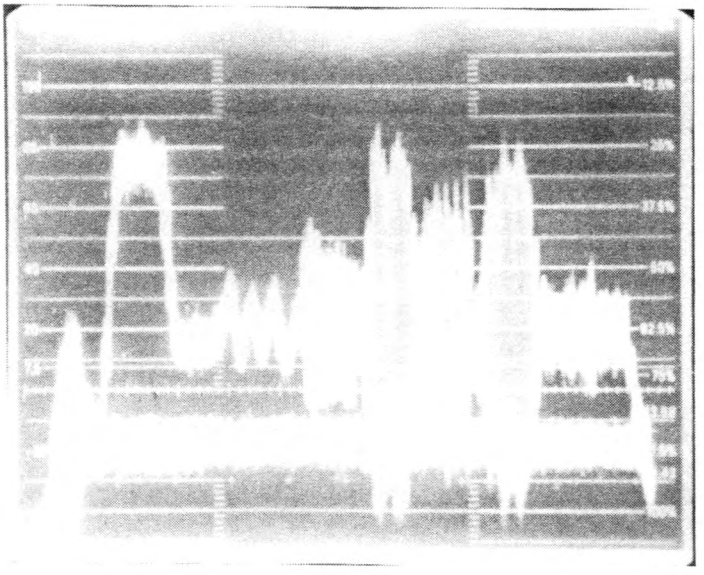

Figure 2D

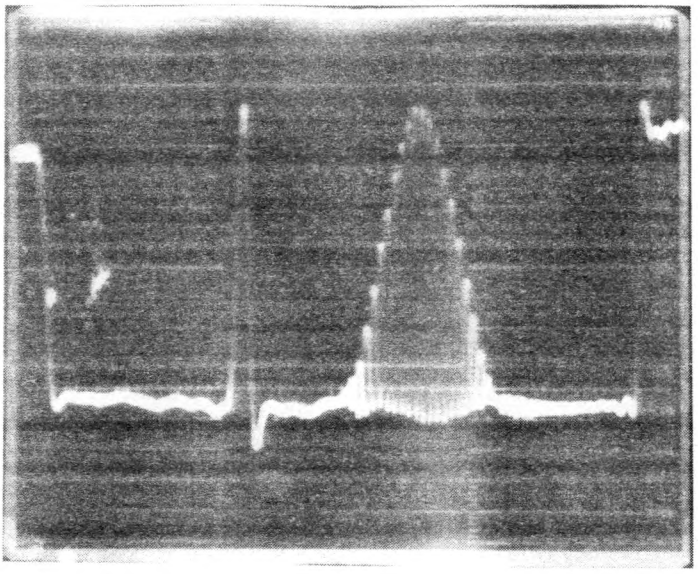

Figure 3A

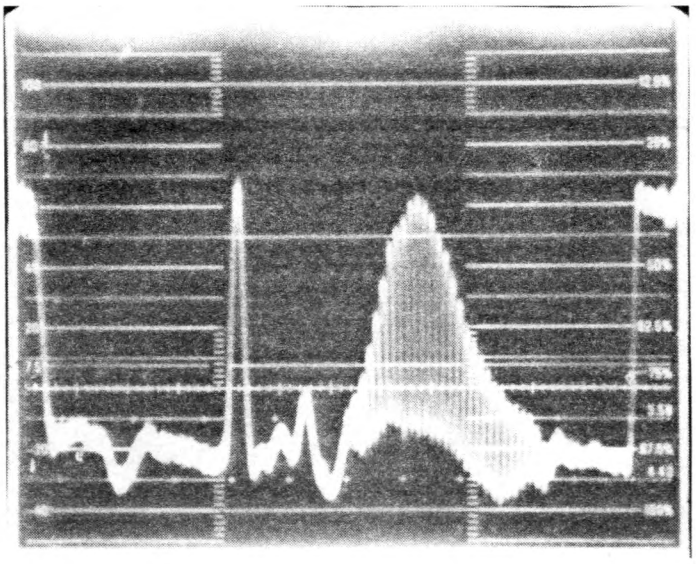

Figure 3B

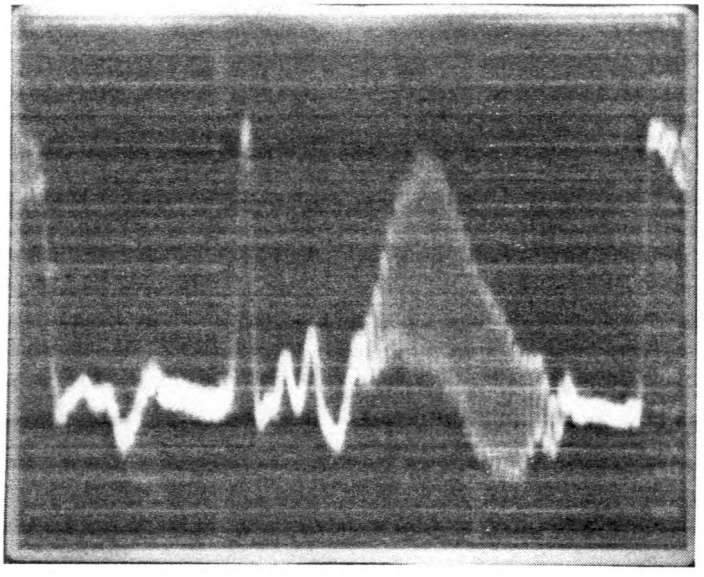

Figure 3C

In Figure 3C, the antenna is oriented such that the two ghosts are more pronounced and the delay is even greater. While degradation is quite perceptible in the picture, many people would still consider this an acceptable picture at normal viewing distances. Figure 3D was taken with the antenna oriented such that the picture is considered unacceptable.

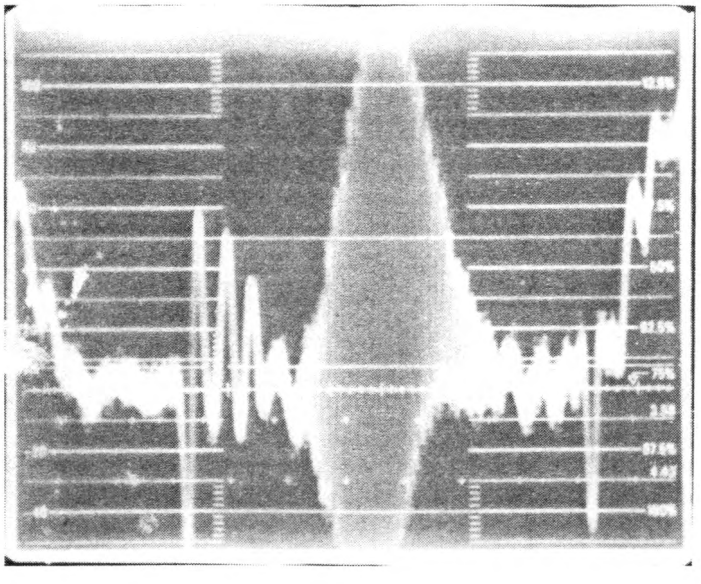

Figure 3D

## Echo Quantification

The above discussion leads us to a method to quantify echo (ghosting), a method which should prove superior to conventional subjective efforts. For example, the first ghost seen in Figures 3B and C is delayed from the main 2T pulse by about 700 ns. A 700 ns delay would create a ghost displaced from the image by about 1% of the screen horizontal width, or about 0.2 inch on a 19-inch screen. We can also measure the relative amplitude of the ghost. This first reflection is about 17 dB below the main signal. The Canadian Department of Communications provides a graph defining the visibiity of an echo as a function of its delay and amplitude.[6] According to this graph, an echo with 700 ns delay is considered visible if its strength is greater than 32 dB below that of the direct signal. Thus, we would expect to see this ghost. The second ghost would be even more visible.

Because of the presence of the 12.5T pulse in the vertical interval, with about 2 microseconds between the two pulses, this technique would be limited to echo delays of less than 2 microseconds (about 0.6 inch on a 19-inch screen). For longer echos, the bar signal which follows the 12.5T pulse can be used. This is shown in Figure 4, where we have an echo delayed about 9 microseconds, and about -25 dB from the direct signal.

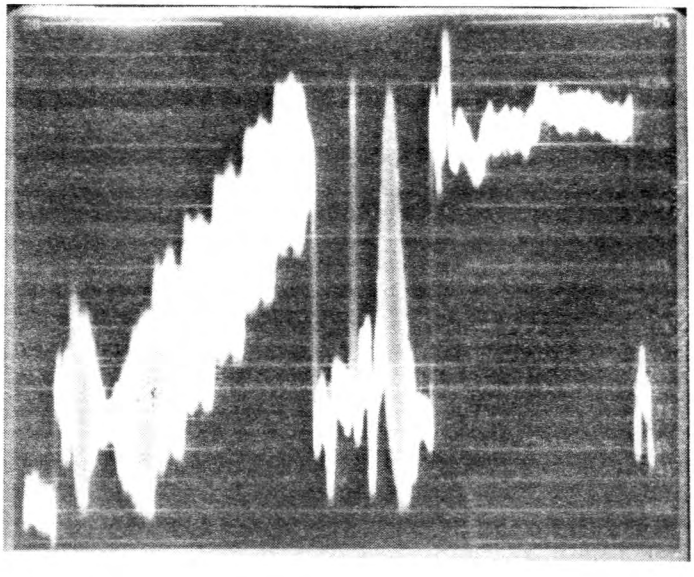

**Figure 4**

In concluding that echo exists, we must first rule out a problem with any other elements in the signal chain (i.e., transmitter, antennas, demodulator). For example, delay errors can cause shifts in the bar signal, that might be confused with ghosting (though delay errors would probably not result in the extremely rapid transition to the final level, as seen in Figure 4).

In all usage of the VITS, the strategy must be to use the variables under control (antenna height and orientation) to seek the best possible performance. The ultimate level of performance achieved will vary considerably from one case to another.

### Check The Headend

Transmitted VITS may also be used to check headend performance on an in-service basis. This can be done by comparing the VITS before and after the headend. This in no way will take the place of the normal proof of performance testing, but it can provide a useful estimate of processor performance, without taking the processor out of service. Remember that we still can only compare the VITS out of the processor with the VITS into the processor. The transmitted VITS will change from time to time, so incoming VITS must be checked before and after each use.

### Conclusion

The above testing must be done with a quality demodulator to be truly meaningful. While the demodulated video out from a TV set having this capabiity may provide an excellent picture, it will not be satisfactory for measurement applications. The measuring techniques discussed are much more sensitive to signal degradation than is the picture. This is as it should be, because of the many sources of degradation between the studio and the home receiver. If any one source of degradation were allowed to introduce noticeable deterioration in the picture, then the overall degradation could rapidly become intolerable.

### REFERENCES BIBLIOGRAPHY

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[3]Straus, T.M. "The Relationship Between NCTA, EIA, and CCIR Definitions of Signal-to-Noise Ratio," IEEE Transactions on Broadcasting, Vol. BC-20, No. 3, Sept. 1974, p. 36ff.

[4]Clayton, L., private correspondence.

[5]Rhodes, C.W. "In-Service Noise Measurements on a CATV System," presented at NCTA Convention, July 6-9, 1971.

[6]Department of Communications, Proof of Performance Procedure for Cable Television Systems, BP24, August 6, 1971, appendix 6.

## SECTION 7
## HEADEND DESIGN

### COMBINING NETWORK 68-14

Model 68-14 is basically a 14 input port combining network. Port 1 has the most attenuation (approximately 21 dB) to its input, while port 14 has the least (approximately 13 dB). The combined output exits at the connector labelled OUTPUT.

An additional feature of model 68-14 is the test input section. If a 9 inch jumper is run from the connector labelled OUTPUT to the connector labelled TEST INPUT, the following are true. First, the SWEEP INPUT port is enabled. Electrically, this is merely an additional input port with approximately .7 dB less input loss than port 14. It has been labelled SWEEP INPUT for convenience to the user who has need for such a connection. Secondly, the -20 dB TEST port is enabled. The signal at this port is -20 dB from the final port labelled TRUNK OUTPUT. With the jumper in place, the TRUNK OUTPUT port is the combined output port.

Below is a block diagram of model 68-14.

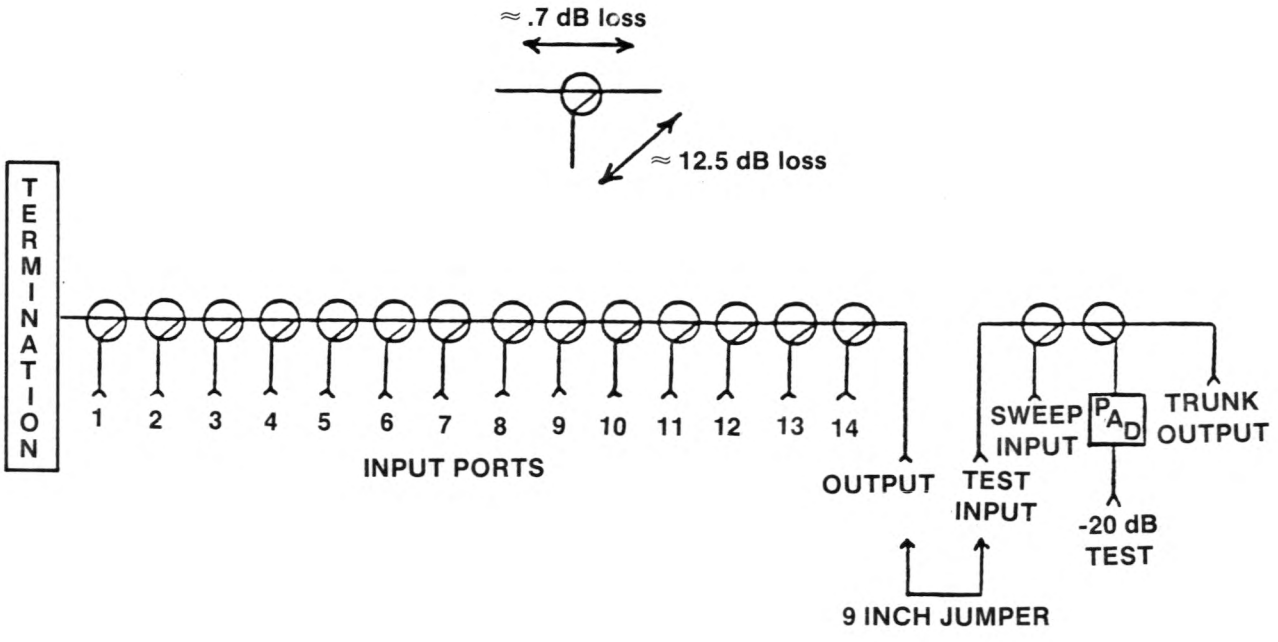



**Figure 7-1. Model 68-14 Block Diagram**

Finally, the four feed-through F connectors on the left of the front panel are another convenience. Connections to these are made by the user if front panel access is required.

| | |
|---|---|
| Frequency Range: | 50-400 MHz |
| Input Return Loss: | 19 dB |
| Output Return Loss: | 16 dB |
| Isolation Between Inputs: | 35 dB |
| Impedance: | 75 ohms |
| Test Point: | -20 dB ± 1 dB |
| Input Loss Port 14: | 13 dB ± 1 dB |
| Input Loss Port 1: | 21 dB ± 2 dB |

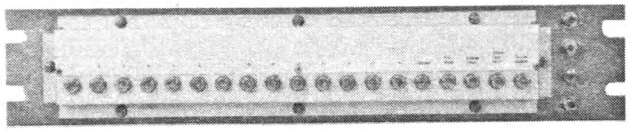


**Figure 7-2. Model 68-14 Rear Panel View**

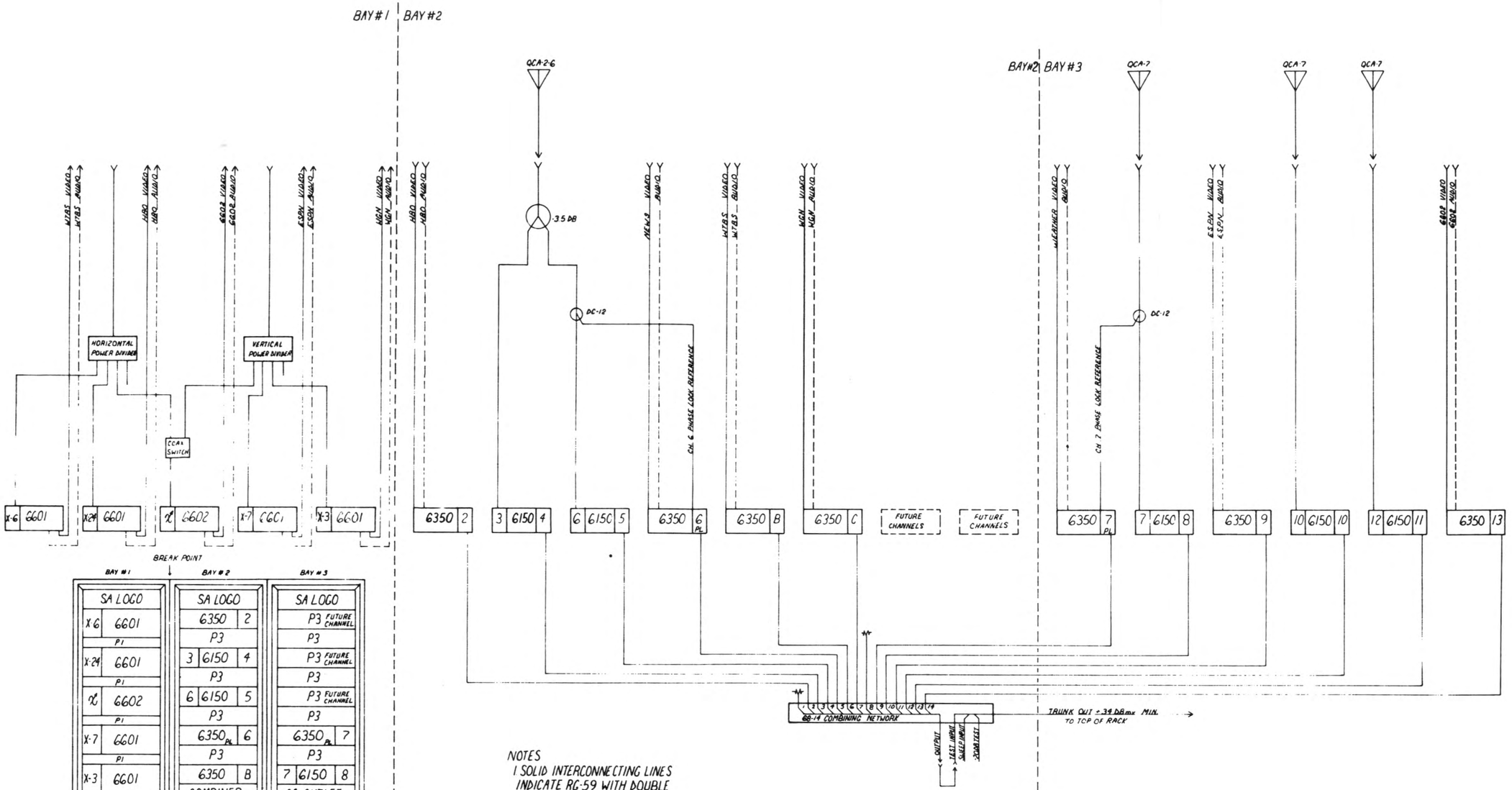



**Figure 7-3. Scope Cable Headend and Earth Station Diagram**

83-HE-80-142

# SECTION 8
# HARMONIC COHERENT HEADENDS

## Theory of Coherent Headends

All amplifiers used in distribution plants inevitably introduce some distortion. This distortion is due to nonlinearities inherent in the active devices used in the amplifiers. During the past several years, new techniques have been developed to drastically reduce the distortion introduced, but it has not been eliminated completely. As cable systems expand in length and as more channels are added, these distortion products become increasingly important. Use of a coherent headend is one method of minimizing the subjective effects of this distortion. While this method doesn't reduce the distortion, it can help to reduce the effect of this distortion as seen by the subscriber.

The voltage transfer function of a CATV distribution plant is often modeled by the equation:

$$E_o = G_1E_i + G_2E_i^2 + G_3E_i3$$

In this expression, $E_o$ is the output voltage, $G_1$ is the overall gain of the plant, and $E_i$ is the input voltage. The second order gain constant, $G_2$ represents second order distortion. Second order distortion produces second harmonics of the input, as well as sum and difference products (a+b, a-b). The third order term, $G_3$, is responsible for the generation of cross modulation (the superimposition of one carrier's modulation on another), gain compression and expansion, intermodulation (generation of products at $2A \pm B$ frequencies), and the so-called triple beats ($a \pm b \pm c$ products).

The purpose of using a coherent headend is to reduce the subjective effects of the beats produced in a distribution system by second- and third-order distortion. For example, consider the effect of the 2A-B third-order intermodulation product involving channels 3 (A) and 2 (B). This 2A-B beat falls at 67.25 MHz, channel 4's picture carrier. If channels 2, 3, and 4 are not all on exactly their nominal frequencies, then this 2A-B product will not fall precisely on channel 4's picture carrier, producing a visible beat. If, on the other hand, all three carriers are at precisely the correct frequency, then the intermodulation product will zero-beat the channel 4 carrier, resulting in no beat being visible on the screen. (Note that the 2A-B product is still present, but it has been forced to coincide in frequency with the desired carrier.) When we state that the beat must fall at precisely the desired carrier frequency, we indeed mean that no frequency error is present. The most practical way to ensure this is by phase locking all carriers to the same master oscillator. If all carriers are forced by a phase-locked loop to hold a strict (but not necessarily

defined) phase relationship to a master oscillator and its harmonics, then the frequencies must be identical to harmonics of the master oscillator (comb generator).

The system described above is an incremental coherent scheme. By this, we mean that all channels (except 5 and 6) fall on their normally assigned carriers, but that the spacing between them is identical, being derived from one oscillator. This coherent scheme has the advantage of retaining standard channels (except for 5 and 6), but has some limitations. These are illustrated by another example.

Consider the triple beat product A + B + C involving channels 2, 3, and 4. This triple beat product falls at 183.75 MHz. This is in the channel 8 passband, 2.5 MHz above channel 8's picture carrier. Incremental coherency cannot force this product to zero beat with the channel 8 carrier. Similarly, the incremental coherent system is not capable of making any of the second order distortion products zero beat. These problems can be overcome with use of a harmonic coherent system. In this system, all picture carriers are at exact multiples of the channel spacing, nominally 6 MHz. In this case, channels 2, 3, 4, and 8 picture carriers are at 54, 60, 66, and 180 MHz, and the A + B + C triple beat between 2, 3, and 4 will fall coherently on channel 8's picture carrier. Note that the only requirement for the comb generator is that the outputs of the comb be exactly at multiples of the channel spacing. This condition is satisfied by deriving all elements of the comb from one oscillator. Should the oscillator drift slightly from 6 MHz, the required condition is still met. Oscillator stability is important only in that all outputs must be within the channel tolerance and within acquisition range of the processor or modulator.

### Harmonic and Incremental Coherent Plans Comparision

In the paragraph "Theory of Coherent Headends", the relative advantages and disadvantages of the two coherent plans were alluded to. This paragraph will attempt to clarify these areas, and will also define the differences between the approaches.

Figure 8-1 graphically compares the two coherent allocation plans with the standard channel allocation plan derived from FCC regulations and standard CATV industry practice. With the incremental coherent allocations, all channels remain at their normal frequencies except for channels 5 and 6. These two channels move up 2 MHz, at the expense of the FM band. This opens up space for an additional channel between 4 and 5, which could be used if the subscriber was furnished with a a suitable converter. Because channels 5 and 6 are offset 2 MHz, some subscribers will not be able to tune them.

All channels are offset in the harmonic coherent assignment; channels 5 and 6 are 0.75 MHz high in frequency, while the remaining channels are 1.25 MHz low. Again, a 6 MHz "hole" is opened between channels 4 and 5. Due to the offset assignments, some subscribers will not be able to tune in the cable channels.

Table 8-1 is a comparision of the important areas of difference between the allocation plans. The offsets from standard channels is indicative of the trouble that some subscribers will have in

tuning the offset channels: all channels are offset in the harmonic plan and channels 5 and 6 are offset in the incremental plan.

When using any nonstandard plan, the possibility exists that a strong local channel can leak into the subscriber's set, causing interference on two cable channels instead of only one. For example, consider the case of a strong local channel 7 leaking into a cable operating with a harmonic coherent headend (figure 8-1 will be of help in visualizing the situation). The interfering carrier will be located 1.25 MHz above the desired carrier, producing a beat on cable channel 7. At the same time, air-channel 7's sound carrier is located .25 MHz below cable-channel 8's picture carrier, producing a beat in cable channel 8. This beat wouldn't be as objectable as would be the beat in cable-channel 7 because the sound carrier is normally transmitted about 6 dB lower in amplitude than the picture carrier. Also, since the undesired signal is below the picture, some roll-off will be experienced due to the Nyquist filter in the TV receiver. However, the beat will still be present, and in extreme cases, one might lose channels 2-for-1 by going to a harmonic frequency assignment. Table 8-1 indicates where these problems might exist.

In a related area, table 8-1 indicates the possibility of phase locking to strong local off-the-air signals to eliminate co-channel beats, and still being able to keep that channel within the comb. Note that, while it is technically feasible to phase lock to one or two off-the-air signals in the incremental plan, equipment to do this is not available at the time of this writing.

Table 8-1 also indicates how much of the FM band is lost to channel 6 in the various allocation plans. This is related to the relative offset of channel 6, and should be considered if FM service on the cable is contemplated. Note also the possibility that a strong FM signal at the low end of the band could leak into the set and create a beat with channel 6, particularly in the incremental allocation.

Finally, table 8-1 designates the particular second- and third-order distortion products that can be made to zero-beat a desired carrier in each coherent plan. Note that, while the harmonic plan can accommodate more types of distortion, the severity of problems caused by each type of distortion tends to improve the position of the incremental plan. This is true using high quality push-pull distribution gear. For example, the number of in-band A + B - C products is much greater than the number of in-band A + B + C products.

### Harmonic Coherent Generation

The incremental coherent plan has been previously discussed before the harmonic plan because it is closer to the standard channel allocation plan. However, in describing the method by which the combs are generated, the two will be dealt with in the opposite order. This is because the concept of the harmonic comb is easier to understand.

The block diagram of the harmonic comb generator is shown in figure 8-2. This block diagram is divided to reflect the actual modules used in the comb generator. Module 1 is the power supply.

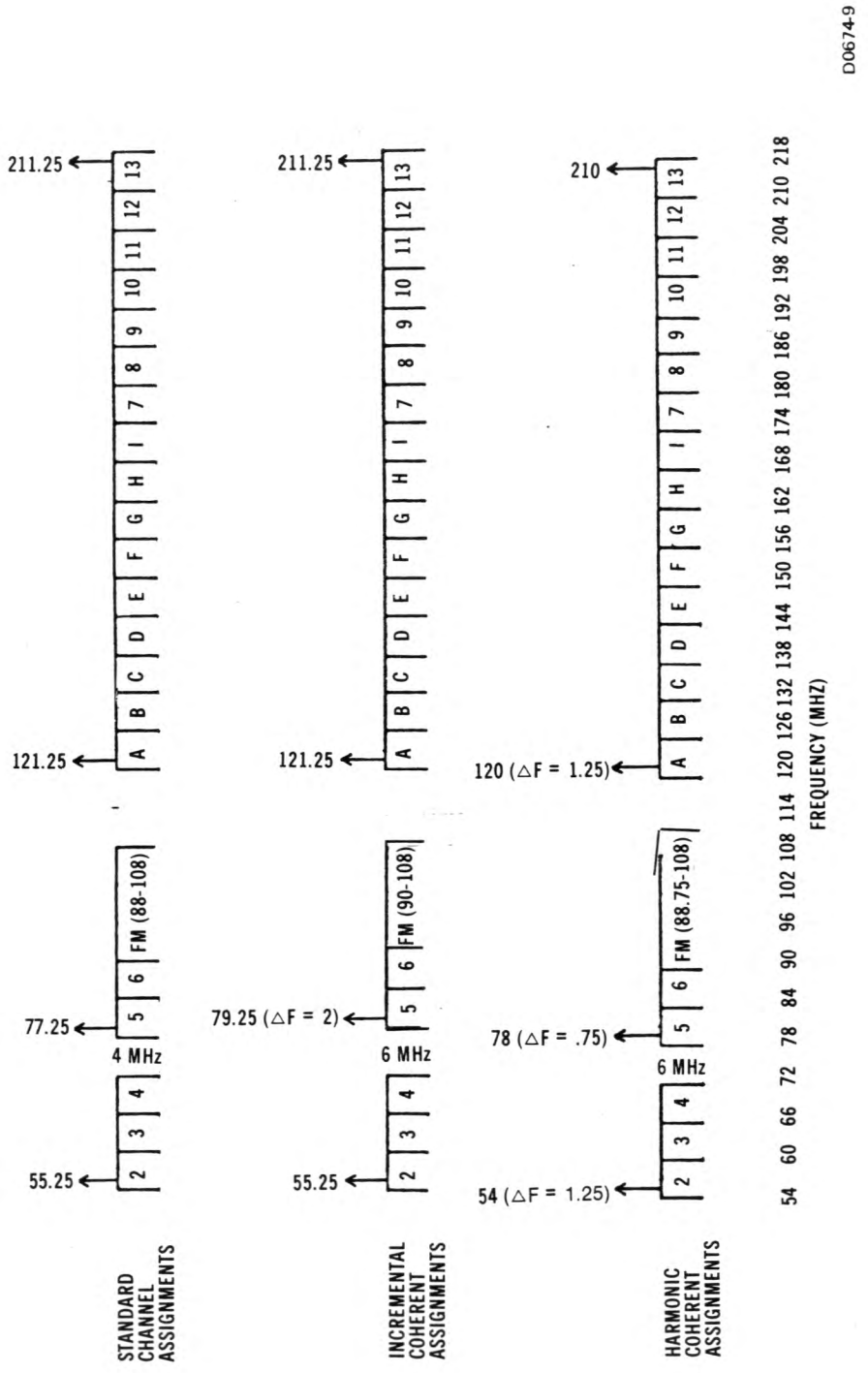



**Figure 8-1. Relative Spectrum of Standard, Incremental and Harmonic Coherent Channels**

## TABLE 8-1. COMPARISON OF ALLOCATION PLANS

| ALLOCATION PLAN / CHARACTERISTIC | STANDARD | INCREMENTAL COHERENT | HARMONIC COHERENT |
|---|---|---|---|
| Channel Assignments (MHz) (Picture Carrier) | Fn = 55.25 + 6n, n=0, 1, 2, . . . except channels 5 and 6 shifted +4 omit n=6–10 | Fn = 55.25 + 6n, n=0, 1, 2, . . . omit n=4, 7–10 | Fn = 6n, n=9, 10, 11, . . . omit n=12, 15–19 |
| Offsets From Standard Channels (MHz) | None | Channels 5, 6: +2 | Channels 2–4, A–13: –1.25<br>Channels 5,6: +0.75 |
| Possibility for Off-Air Channel to Overlap Two Cable Channels | N/A | Channel 6 only | All standard FCC channels except 4, 6, 13 |
| Spectrum Lost From Low End of FM Band | None | 2 MHz | 0.75 MHz |
| Coherent Headend Can Zero Beat The Following Distortion Products: | | | |
| 2nd Harmonic | | | ✓ |
| A+B | | | ✓ |
| A–B | | | ✓ |
| 2A+B | | | ✓ |
| 2A–B | | ✓ | ✓ |
| A+B+C | | | ✓ |
| A+B–C | | ✓ | ✓ |
| A–B–C | | | ✓ |
| 3rd Harmonic | | | ✓ |
| Ability to Lock to Strong Local Signals | Unlimited | 1 or 2 channels* (except channels 5, 6) | None |

*Equipment not available as of this writing.

Module 2 contains a 6-MHz oscillator, comb generator, and lowpass filter which limits the maximum comb frequency to just over 400 MHz. Module 3 contains a hybrid amplifier, which raises the level of the comb to one adequate for driving all processors and modulators. A hybrid four-way splitter is included on the comb generator chassis, to provide four identical outputs. The comb is then further split externally to provide a signal for each processor or modulator.

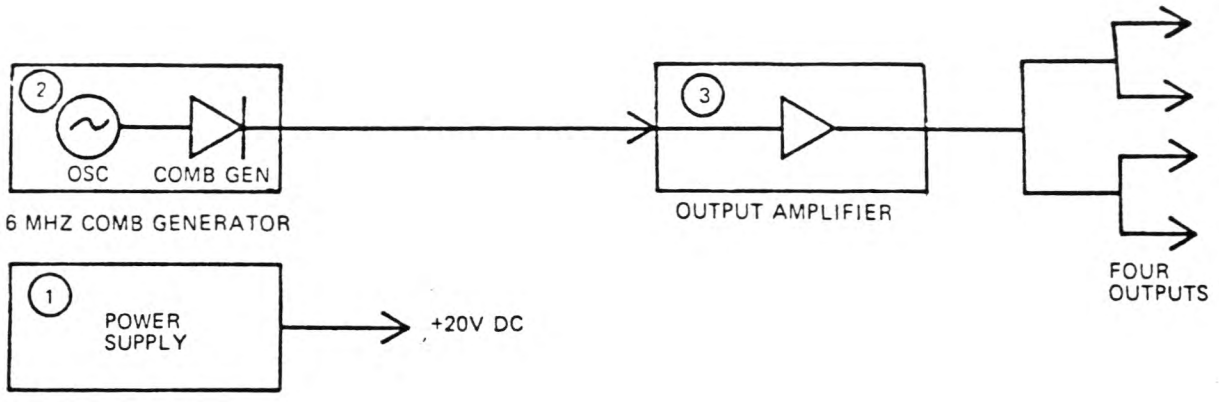



**Figure 8-2. Block Diagram, Harmonic Comb Generator**

To generate a harmonic comb, all that is required is an oscillator running at the frequency equal to the desired spacing between elements of the comb. Harmonics of the oscillator then form the comb. In order to provide a fairly uniform comb from 6 to 400 MHz, one must intentionally generate controlled harmonics, rather than to simply rely on the normal high harmonic content of the oscillator output. From the study of Fourrier transforms, we know that one waveform that is rich in harmonics is a rectangular wave such as is shown in figure 8-3.

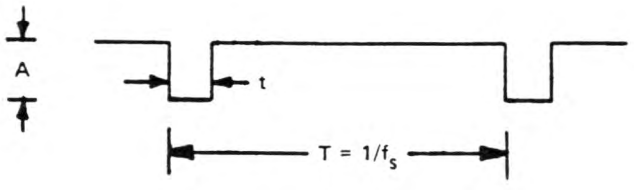
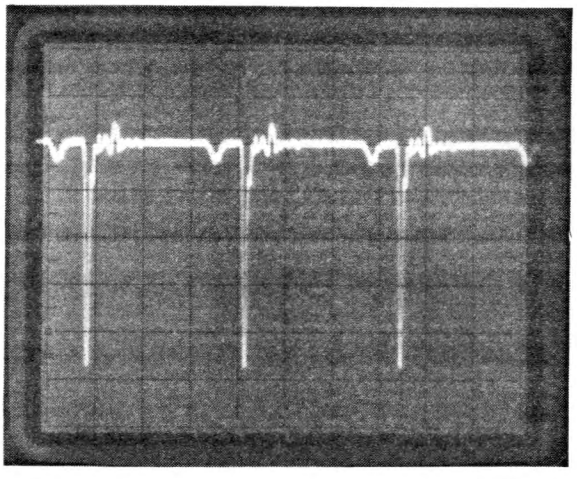



**Figure 8-3. Waveform Used to Generate Basic Comb**

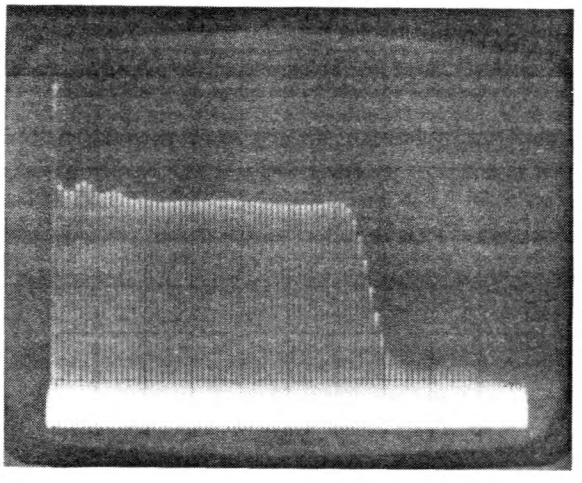
The time between successive pulses, T, is equal to the reciprocal of the spacing between elements of the comb, $f_s$. Now suppose that we begin modifying this waveform by repeatedly halving the width of the pulse, t. Each time that we halve the width, we double its amplitude, A. In the limit, as we approach zero pulse width and infinite amplitude, we approach an impulse which, in the frequency domain, possesses equal energy at each harmonic of $f_s$ from $f_s$ to infinite frequency. This is a limiting condition which cannot actually be reached. Fortunately, the comb need not extend to infinite frequency. Figure 8-4 shows the time and frequency domain outputs from the comb generator module.

.2V/DIV VERT
50 NS/DIV HORIZ
0 DB PAD IN MODULE



a) TIME DOMAIN

GAIN REF: +14 DBM
HORIZ: 50 MHZ/DIV
0 DB PAD IN MODULE



b) FREQUENCY DOMAIN

**Figure 8-4. Comb Generator Output**

These two waveforms are respectively displayed on a high frequency oscilloscope and on a spectrum analyzer. These photographs were taken at the output of the comb generator module, after the comb has been intentionally bandpass limited to just over 300 MHz. The lowpass filter used to limit the upper frequencies is partially responsible for the deviation of the time domain waveform from the theoretical waveform shown in figure 8-3. Without this intentional filtering, the output would appear more like figure 8-3, and the spectrum would extend to about 800 or 1000 MHz, with the amplitude varying with frequency. The output of the comb is intentionally limited to about 400 MHz to avoid the possibility of higher frequency components leaking out of the comb distributions wiring and causing interference with a UHF input or other signal.

### Incremental Comb Generation

The incremental comb generator is similar to the harmonic comb generator in that the spacing between channels is set by a 6-MHz oscillator. However, the output frequencies are not exact multiples of 6 MHz; rather, they start at 55.25 MHz and progress upward. This is accompanied by modulating the output of a translation oscillator by the comb (figure 8-5).

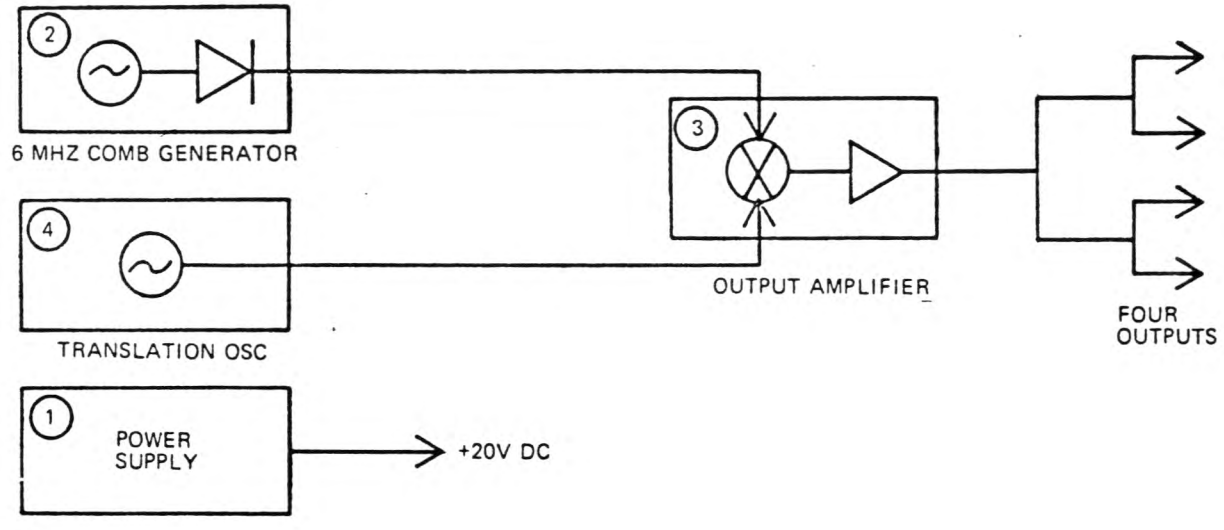
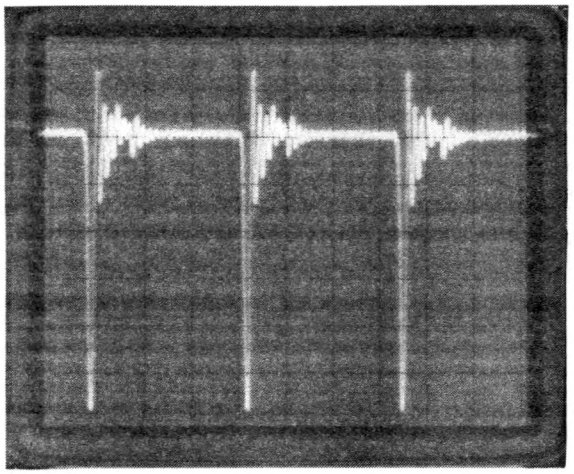



**Figure 8-5. Incremental Comb Generator Block Diagram**

Normally, the translation oscillator operates at 175.25 MHz, channel 7's picture carrier. The output of the comb generator is limited to less than 150 MHz in this case, to avoid sideband foldover at zero frequency. Figure 8-6 shows the output of the incremental comb generator (module 2 in Figure 8-5) in both the time and the frequency domains. To the frequency domain display has been added the translation oscillator output to show the relative frequencies involved.

As shown in figure 8-5, the outputs from the comb and the translation oscillator are mixed together in the output amplifier. The comb modulates the translation oscillator output, producing a double sideband spectrum around the translation oscillator frequency. A double balanced mixer is employed, with provision to unbalance it to a selected degree with respect to the translation oscillator input. This allows the translation oscillator to be inserted into the output comb at the proper level. Figure 8-7 shows the time and frequency domain outputs from the output amplifier. Because of the shift in the starting frequency from 6 MHz, the time domain display cannot be held still, and the actual display seen will appear similar to figure 8-7a.

VERT: .2 V/DIV
HORIZ: 50 NS/DIV
0 DB PAD IN MODULE



GAIN REF: +10 DBM
HORIZ: 20 MHZ/DIV
0 DB PAD IN MODULE

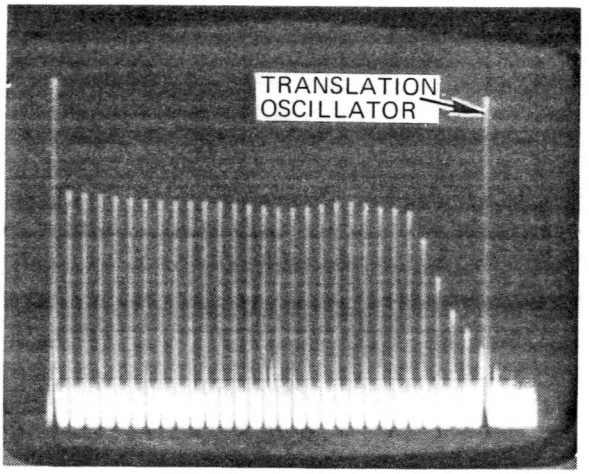
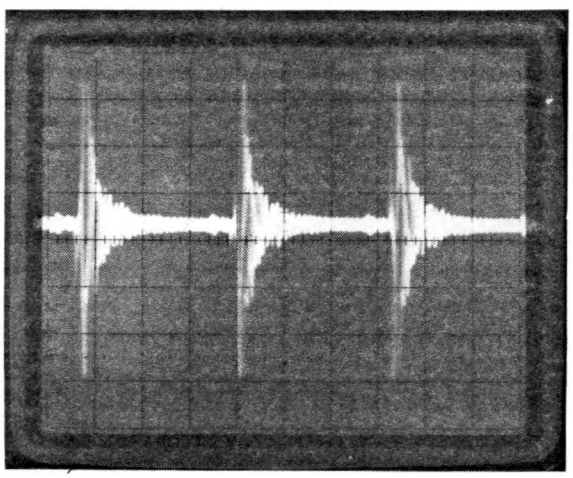


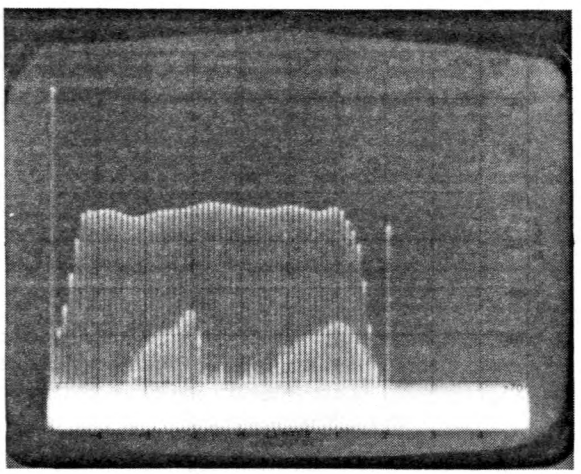

**Figure 8-6. Incremental Comb Generator Output**

VERT: 0.2 V/DIV
HORIZ: 50 NS/DIV
12 DB PAD IN CG MODULE



A

GAIN REF: +10 DBM
HORIZ: 50 MHZ/DIV
12 DB PAD IN CG MODULE



B

**Figure 8-7. Output of Incremental Comb Generator**

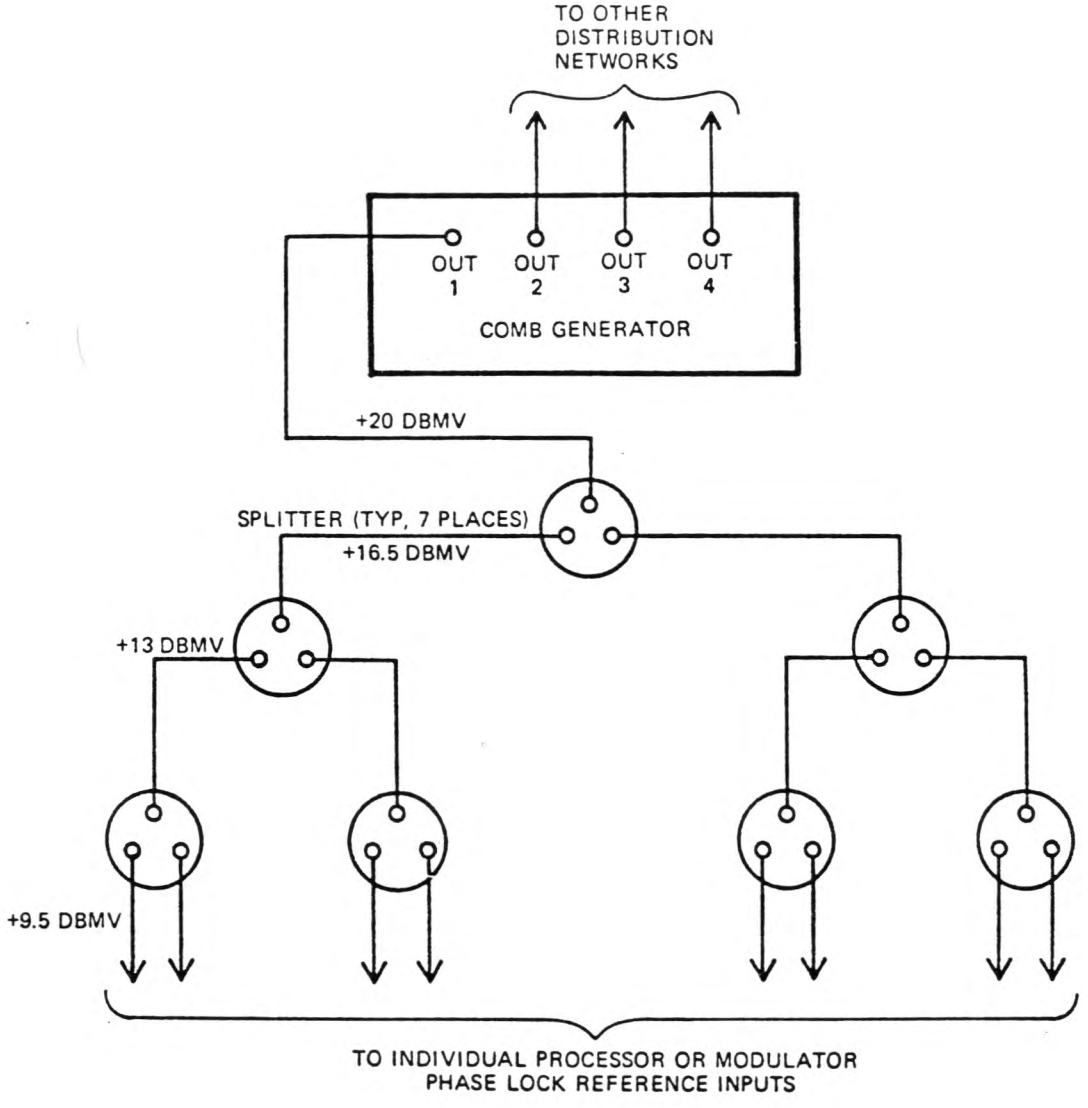



**Figure 8-8. Typical Distribution Schematic**

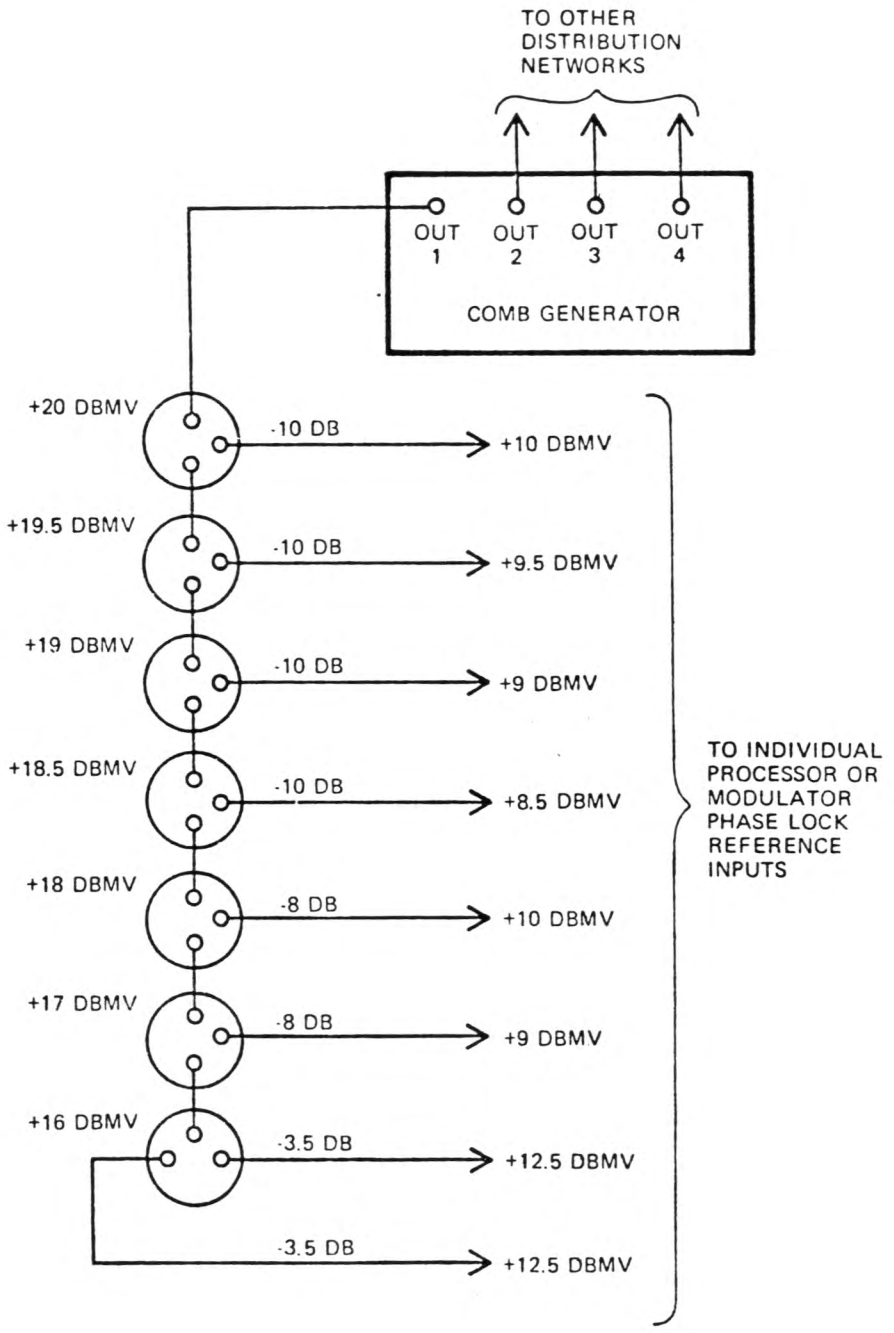



**Figure 8-9. Typical Rectangular Distribution Schematic**

## SECTION 9
## HEADEND SIGNAL SWITCHING

**Reprint of**

# HEADEND SIGNAL SWITCHING TECHNIQUES*

**by**
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### Introduction

Signal switching has been practiced in CATV headends as a means of achieving better utilization of a limited number of channels and for commercial deletion and non duplication substitution. In designing a headend switching system, several problems need to be addressed. A selection can be made between baseband and IF switching, each having advantages under certain conditions. Different types of hardware are available to implement switching operations, including mechanical switches and relays, and electronic switches. When multiple inputs or outputs are involved in a switching operation, differing methods may be employed to implement signal distribution, with the method selected affecting isolation, loss, and complexity.

Several methods exist for actuating a switching operation, ranging from manual to automatic detection of program coincidence, to activation upon loss of one input. A number of case histories are examined to illustrate differing switching designs.

### Baseband vs IF Switching

The large headends being designed and installed in the 1980's often contain upwards of 50 channels of output. In these headends, where there is a low ratio of off air to video program sources, cable operators often demodulate the off air signals and then remodulate them. This technique, called "Demod-Remod" offers several advantages in signal switching which are discussed later in this section. Demod-remod headends are still in the minority, however, and so most of this paper discusses IF switching.

While IF switching is frequently attractive, it also offers some drawbacks. Probably the most severe of these is the need for substantial isolation. When a number of cables carrying different signals at nearly the same carrier frequency are in close proximity to each other, signal leakage can cause severe problems. These problems are exacerbated by the limited isolation available in switches and patch panels. Since the frequencies used are high compared to baseband frequencies, isolation is more difficult. Isolation problems may be reduced by choosing switches with more than sufficient isolation at the IF frequency and by using cable with a high percentage of shielding (double shielded cable is frequently chosen). Naturally, connectors must be installed with care to ensure long term integrity of the shield connection. Cables should be separated as much as practicable and if possible the switching matrix should be distributed through the headend. In fact, a preferred technique when possible, is to locate most or all switching within the processor or modulator, thus minimizing external cabling.

*Presented at: Canadian Cable Television Association, 22nd Annual Convention, Trade Show & Videotheque, April 2 - 5, 1979, Toronto.

An alternative to IF switching is baseband switching, long popular in broadcast practice. Equipment is available for baseband switching, but the total cost may be higher if signals must be demodulated for switching. Too, extra demodulation and modulation processes can introduce additional signal distortion.

On the other hand, baseband switching offers benefits which sometimes outweigh the disadvantages. Since switching is accomplished at lower frequencies, isolation problems are potentially less severe. Interface levels are standardized at 1 volt p-p, eliminating in many cases the need for amplifiers or attenuators. Matrix switching is less complex since it is practical to bridge a number of high impedance video amplifiers across the input. A frequently useful technique, available commercially, is detection of loss of video and subsequent activation of a switch. Finally, operating at baseband can offer the opportunity to process the signal to restore deficient amplitude, sync, or burst.

**Table 1. Comparison of Baseband vs IF Switching**

| Factor | Baseband Switching | IF Switching |
|---|---|---|
| Interface Levels | Standardized | Not Standardized |
| Isolation Problems | Less Severe | More Severe |
| Cabling Complexity (Large Matrices) | Less Severe | More Severe |
| Cost and Non-Linear Distortion | Depends upon whether baseband or IF signals are available. | |

Table 1 compares baseband switching to IF switching. Since most CATV switching is done at IF, this type of switching is emphasized in the paper.

## Switching Hardware

When selecting switching hardware, several options are available. The simplest is a coaxial switch, usually of the single pole double throw variety. The cost is relatively low, power is not required, so the switch is not affected by power supply failure. The switch may be used for both baseband and IF switching. However, a simple switch cannot be remotely controlled (as from a programmer), takes up considerable panel space, and makes cabling difficult since mounting position is restricted to a configuration which makes it easy for the operator to reach the switch. All mechanical switches exhibit wearout failure after some number of cycles, and mechanical contacts have long been known for susceptibiity to failure.

Coaxial relays are frequently used for switching. Like switches, they can be used either at IF or baseband, though with the lower cost units isolation at IF may be a problem. Relays are controlled remotely and consume power when actuated. If power should fail, they revert to the non-actuated state. (Magnetic latching relays will retain either state upon loss of power, but they are expensive.) Since relays are remotely controlled, they may be placed where needed, frequently simplifying coaxial cable routing and possibly easing isolation problems. On the other

hand, coaxial relays exhibit a limited number of activation cycles and can again experience mechanical contact failure. If it is desired to have a port terminated when not connected, a relatively large cost differential can result.

Solid state relays (switches) overcome the problems of limited number of activation cycles and mechanical failure problems. The range of frequencies over which a solid state relay operates effectively is somewhat restricted, though they are available for use at either baseband or IF. Solid state relays consume power in both states (though less than is consumed by an activated mechanical relay), and are susceptible to power failure in either state. However, they offer excellent isolation at reasonable cost. Termination of the unused port is standard. As opposed to mechanical relays, the power required to activate them is compatible with integrated circuit logic. Figure 1 is a photograph of a typical mechanical switch, a coaxial relay, and a solid state relay.

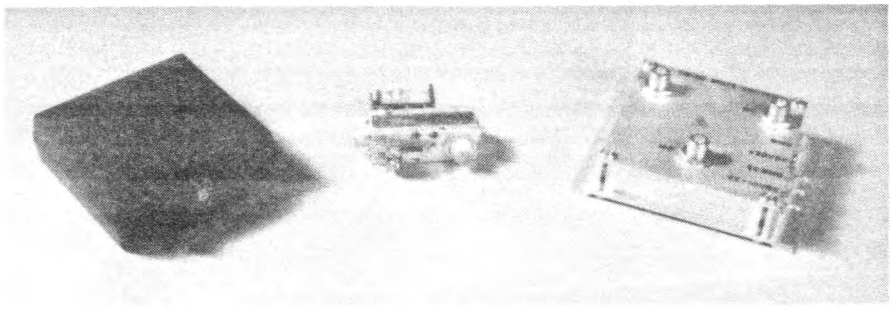


**Figure 1. (L to R) Mechanical Switch Coaxial Relay, Solid State Relay**

Patch panels are often used in manual switching systems. They are simple and flexible, though they may not be cost effective if the flexibility isn't needed. Patch panels may be ordered with built in termination of unused ports, or with switching to an alternate source if a jumper is unplugged. As with coaxial switches, the cable routing must accommodate the panel, not the other way around.

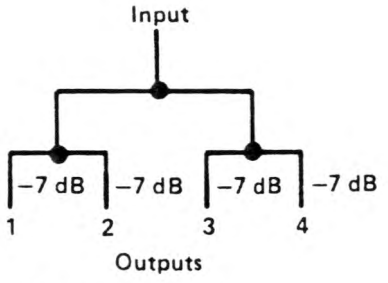
### Basic Splitting Concepts

When an IF signal must be simultaneously supplied to a number of switches, it must of course pass through a number of splitters. (This is an advantage of baseband switching — — the splitters could be replaced with high impedance input amplifiers, without suffering a power loss.) Each time a signal is split two ways, a minimum of 3 dB is lost at each output since half the power must be routed in each direction. In practice, the loss is a bit higher due to losses in the splitter. Signals can be amplified if required, but if the signal level drops too low amplification will add noise. Considering the practical maximum carrier-to-noise ratio found in headends (around 60 dB) and the noise figure of many typical IF amplifiers, a good rule to thumb might be to maintain

levels of at least +10 dBmV, and preferably +15 to +20 dBmV. In order to reduce crosstalk problems, all signals in a common area should be at about the same level.

A splitting arrangement may be configured either of two ways: as a tree or in a rectangular fashion. The two are illustrated in Figure 2. which shows one input and four outputs.

**Figure 2a**




A) Tree Distribution

**Figure 2b**

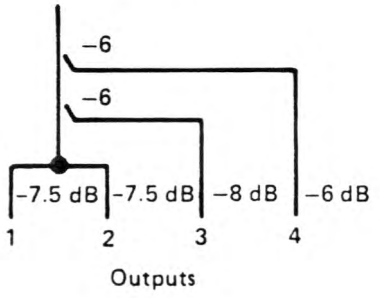



**Figure 2. Basic Multiple Splitting Configurations**

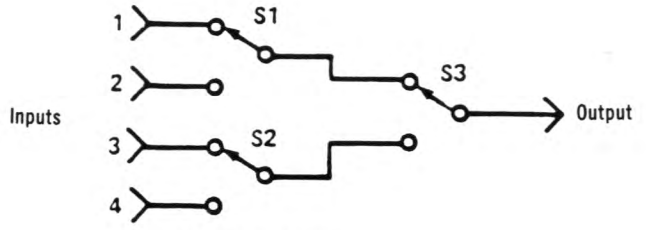
The tree configuration offers commonality of components and equal levels at all outputs (but only if the number of outputs is equal to $2^n$ where n is an integer). Physically, a tree distribution structure offers convenient wiring if the signals disperse in different directions after the splitting point (as with signals that are cabled to different racks).

The rectangular distribution shown in Figure 2b uses a combination of directional couplers and splitters. In general, the signal levels at the different ports will not be consistent regardless of the number of outputs. This type of distribution would be easier to cable if most outputs are in one direction from the common point. For example, if one signal had to be routed to each processor in a rack, a rectangular distribution arrangement with a directional coupler near each processor would result in the simplest wiring. Note also that a majority of CATV distribution plant is thusly fashioned.

## Basic Multiple Switching Configurations

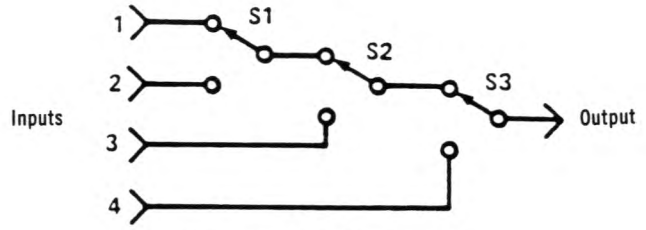
In a manner similar to that shown above for signal distribution, one can envision two ways of configuring switching to permit a number of IF sources to be supplied to one load. These are shown in Figure 3. (Consistent with common practice, the term switch will be used to mean switch or relay unless ambiguity would result from such usage.)

**Figure 3a**




A) Tree Switching

**Figure 3b**




B) Rectangular Switching

**Figure 3. Basic Switching Configurations**

The tree switching configuration places the minimum number of switches in the longest signal path, which might be a consideration if switch insertion loss is high. On the other hand, some logic is required to activate multiple switches in order to select a given input. For example, with the configuration shown in Figure 3a, selection of input 4 would require both switches S2 and S3 to be activated. If input selection is manual, then the operator will have to throw the switches to set up this input. In a large switching operation, this would certainly cause great confusion, resulting in frequent errors. Alternatively, logic could be inserted if mechanical or solid state relays were used, to permit several relays to be actuated from one command. This would simplify command generation, but at the expense of some additional hardware.

The rectangular matrix shown in Figure 3b results in only one switch (or relay) being actuated at any time. This should reduce operator errors without requiring any driver logic. However, since more switches are in some signal paths (e.g., from input 1), switch insertion loss may be more of a factor.

An important consideration if switch isolation is marginal, is the overall isolation of signal sources. Suppose input 1 is selected. In the rectangular configuration signal leakage from inputs not selected will be reduced from the selected signal by the isolation of one switch. On the other hand, with the tree arrangement of Figure 3a, the signal on inputs 2 and 3 are reduced by the isolation of one switch, but the signal on input 4 is reduced by the isolation of two switches. With the configurations shown and assuming equal input signal levels on all inputs, a simple calculation (see appendix 1) will show the relative level of undesired signals on the output. For example, if all switches have an isolation of 60 dB, then the total power of all undesired carriers in the output is 57 dB below the desired carrier with the tree arrangement. With the rectangular arrangement the undesired signals are only 55 dB below the desired signal.

Of course, switch insertion loss would further degrade the net isolation. With provision for switching more inputs, the reduced isolation of the rectangular matrix would be more dramatic.

## Single Processor Switching

In this section we show several simple switching arrangements which can be accomplished generally within the confines of a single processor. Switching in this context is used to place one of two IF signals on a single channel.

### 1. Low Level IF-External Command

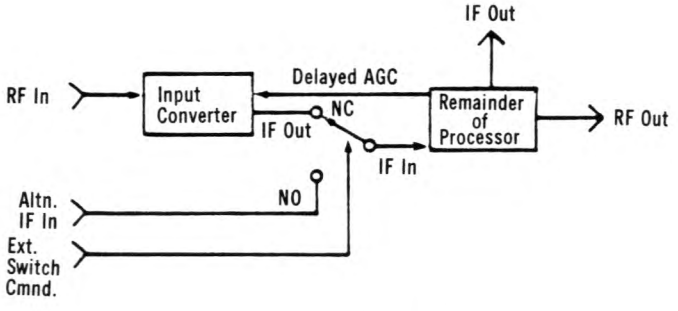



**Figure 4. Low Level IF Switching**

In this basic switching operation, the processor signal path is broken just after the input converter and a switch is inserted. (Figure 4). This switch is normally closed permitting the processor to operate in the usual manner. However, when it is activated from an external command, a second IF signal, from an external source, is switched to the IF. The external command is frequently from an automatic switch controller and is usually in the form of a contact closure, though a voltage level may sometimes be used.

Notice that all IF signal processing (filtering, AGC and adjustment of sound carrier level) will be applied to both the normal and alternate IF signals. Also, if a high level IF output is provided for sourcing additional equipment, the selected program will appear on it. Since only one switch is used, one might expect the isolation requirements of the switch to be equal to the amount by which it is desired to have spurious signals suppressed (usually 60 dB in contemporary CATV practice). This is true if no other sources of interference are anticipated and if both signals are at equal levels. If one signal is higher in level than the other, then this difference increases the isolation requirement of the switch.

Generally, a portion of the AGC range over which a processor operates is contributed by delayed AGC in the input converter. This permits the input circuitry to operate at maximum gain for low level input signals, permitting improved carrier-to-noise ratio as the input signal amplitude rises. Then after a certain input level, the RF gain is reduced as the incoming signal level increases. This is done to prevent mixer overload. Care should be exercised to ensure that the maximum signal level applied to the external IF input doesn't exceed that supplied normally from the input converter (the level from the input converter being limited by delayed AGC). This will ensure that the IF section of the processor isn't overloaded.

**2. High Level IF Switching**

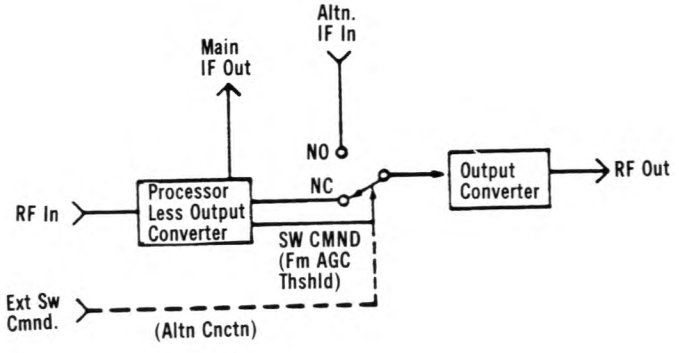



**Figure 5. High Level If Switching**

In this arrangement, the IF path is broken just before the output converter, after filtering and AGC. The output level control may be on either side of the switch depending upon the configuration of the processor. Since the alternate IF signal will not receive processing, all bandpass filtering, AGC action and adjustment of the sound carrier level must take place externally. Often an auxiliary IF AGC module is available which can be placed in the alternate IF input line, and which offers limited range AGC action. Since AGC is frequently not applied to the IF signal as it enters the output converter (though this could be done if required), the main and alternate IF signals at the switch must of necessity be equal in level. Thus, the isolation requirement of the switch will be equal to the amount by which it is desired to suppress the signal not selected.

Note from Figure 5 that placement of the switch after the AGC circuitry permits an alternate method of activation. A threshold detector, usually part of the standby carrier circuitry, may activate when the main RF signal drops below a certain level (as when the station leaves the air). Rather than turning on the standby carrier, the detector can be wired to activate the switch. Thus, an alternate program may be automatically selected when the primary station leaves the air. Switch activation from an external command is also possible, and activation from external command *or* when the station leaves the air is entirely practicable. When the switch is thrown from an external command, the main program IF signal is still available for routing to other equipment.

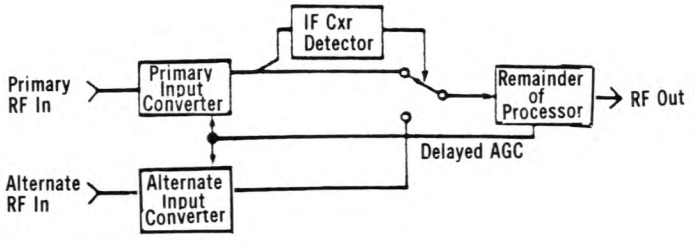



**Figure 6. Low Level Automatic Switching**

### 3. Low Level Switching Upon Loss of Main RF

This configuration presents a low cost way to switch to an alternate RF signal when the main RF signal is lost. Essentially the same as the low level switching shown above, the alternate IF signal is frequently developed from a second input converter. One would favor this arrangement when the desired alternate program is not otherwise available at IF. The IF detector samples the IF from the primary input converter and activates the IF switch upon loss of the primary signal.

Note that delayed AGC must be supplied to both input converters. Also, since the two RF input signals will not generally be at the same level, the switch isolation requirements will exceed the required suppression of the non-selected signal. Depending upon the equipment selected, the functions shown will be accommodated in one or more chassis.

## Examples Involving Baseband Switching

Examples in this section are chosen to illustrate simple switching involving baseband video signals. Generally these configurations involve modulators, but one configuration involving only a processor is shown.

## 1. Switching Upon Loss of Main Video

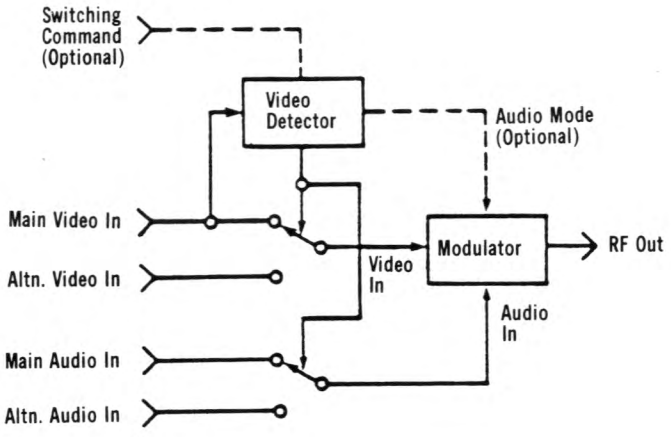



**Figure 7. Switching Upon Loss of Main Video**

A video switch may be teamed with a video detector to permit switching from one video source to another upon loss of the first source. (Often the combination of the switch and detector is simply referred to as a video switch.) For example, suppose that it was desired to share a channel between a message service and programming derived from a video tape recorder (VTR). The VTR could be supplied to the main video input and the message video to the alternate input. Then when a video tape was to be played, all that would be necessary is for the tape to be turned on. Receipt of video would activate the switch, and loss of video at the end of the tape would result in transfer back to video. Note the inclusion of a switch for audio, which is driven by the video detector to permit audio-follow-video action. In some instances one video source is accompanied by baseband audio and the other by a 4.5 MHz sound subcarrier. Some modulators may be programmed to accept either, by a voltage applied to a programming input. In this case the video switch controls the sound mode on the modulator and the audio switch is unnecessary. When both programs are accompanied by the sound subcarrier, audio switching is inherent since the subcarrier is taken after the video switch.

When performing switching, one needs to be certain that video levels are indeed all standardized (normally to one volt peak-to-peak). If this cannot be assured, then one may wish to insert a video AGC circuit between the video switch and the modulator. An AGC circuit inserted at this point should only normalize the sync tip level and should not operate on active video. Similarly, one may consider using AGC on the audio, though the selection of processing equipment should be made considering the consequences of eliminating desired dynamic range.

## 2. Video Switch Controls IF Switch

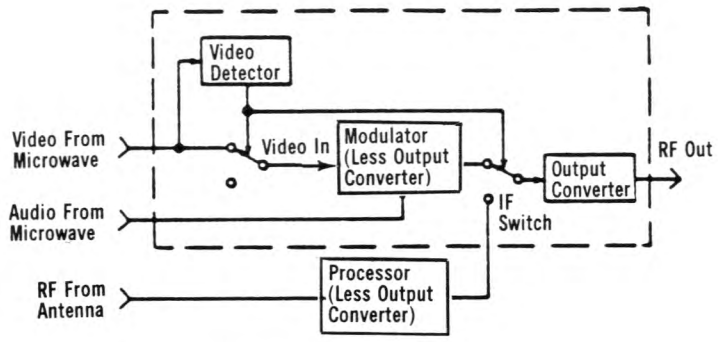



**Figure 8. IF Switching Upon Loss of Video**

This application is shown in the context of solving a particular problem. A short microwave link is used to deliver high quality signals from a distant station. The receive site is close enough to the original transmitter to pick up a signal, although it is too weak to produce satisfactory pictures. It is desired to use the off-air signal however, as a back-up to the microwave delivered signal. A video switch is used as before, but rather than using an input to the alternate video, the video detector drives an IF switch. If the microwave link fails, the video switch activates the IF switch, which receives an IF signal from a processor. The processor is receiving the signal off-air from a television antenna. Thus, if the microwave link fails a back-up signal is automatically inserted.

An additional consideration in the selection of the video detector should be mentioned in this context. A simple video detector could be constructed by clamping the sync tip to a fixed voltage and peak detecting the signal. A threshold could be established low enough to avoid triggering on a black signal. However, such a circuit would be susceptible to falsely assuming a signal was present if noise existed at its input. This would be the case, for example, if the microwave transmitter failed, resulting in a high amplitude noise signal being supplied from the microwave receiver. An improved technique of detecting video is to make use of the fact that a video signal has a high concentration of energy around 15.75 kHz. A bandpass filter at this frequency is inserted in the video path, and the output of this filter is detected to provide the switching signal. Then if the signal is replaced by broadband noise, the energy within the passband is greatly reduced, and accurate determination of signal loss is achieved.

## 3. Processor Switching Upon Loss of Modulation

Occasionally the requirement has arisen that an incoming signal to a processor will always be present, but may not always be modulated. It is desired to switch to an alternate IF signal when the modulation is removed. For example, this was requested recently by one CATV system which received a signal via supertrunk from a neighboring CATV system. The signal was an imported

station delivered to the originating system via microwave. At times the microwave system failed, removing modulation from the carrier on the supertrunk. However, since the video was applied to a modulator, the carrier remained on. Upon microwave signal loss, the receiving system wanted to substitute the IF signal from another processor. Several configurations are possible to solve this problem, with the most economical being shown in Figure 9.

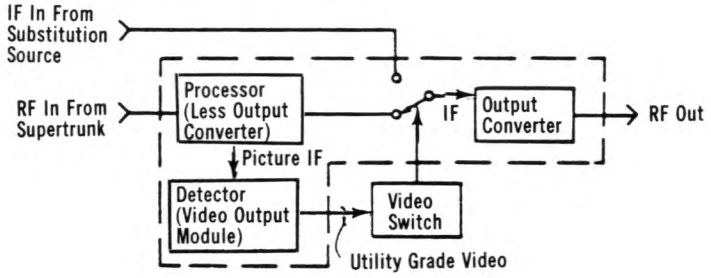



**Figure 9. Processor Switching Upon Loss of Modulation**

A sample of the video IF signal is taken from a directional coupler at a point just before the AGC detector. The signal at this point doesn't include a sound carrier, and happens to be at a suitable level that a sample taken from an 18 dB directional coupler can be supplied to a demodulator video output module. This module includes an envelope detector, and in this configuration will provide a utility grade video output. Note that, due to lack of a Nyquist slope filter and proper receiver group delay characteristics, the video is not suitable for uses other than the one shown.

The utility grade video signal is then supplied to a video switch module. The video presence detector in this switch in turn drives an IF switch, which connects the alternate IF source when modulation is lost on the incoming signal. As shown, the IF switching is accomplished as in Figure 5. If low level IF switching as in Figure 4 is desired, then the video would have to be derived from an external demodulator. Only the input converter could be shared between the processor and the demodulator. In the case that inspired this example, the customer used a complete envelope demodulator and split the incoming RF signal between the processor and demodulator. While this was not the minimum cost configuration, it did provide for minimum special cabling, and was a vehicle for acquiring spare modules.

## More Complex Examples

The concepts covered above may be expanded to permit more complex switching arrangements. Examples shown are either taken directly from actual designs or are modified to illustrate possible applications involving more than one type of switching.

### 1. Off-Air Multiple Backup

In this case, an operator had a public access channel supplied by a modulator, which was in use only sporadically. When public access was not in use, sync was not supplied to the modulator. At

such times, the operator wanted to place off-air channels 3 or 18 on the channel. Selection of which channel was to be placed on the cable is based on which channels are on the air, channel 3 having higher priority. Both stations were picked up off the air, but were not otherwise carried on the cable. All switching was to be automatic.

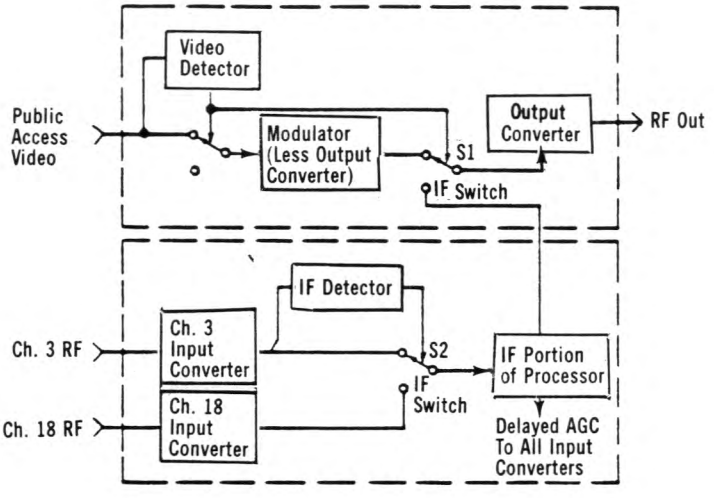



**Figure 10. Priority Switching**

The public access video is supplied to a video switch used only as a video detector, then to the modulator. High level IF switch S1 before the modulator output converter, is driven from the video switch. If the video is removed, S1 is thrown, accepting an IF signal from S2 (via the IF portion of a processor). If channel 3 is on the air, its IF signal is supplied to the output converter. Should channel 3 be off the air, S2 is thrown by the IF detector, to accept an input from the Ch. 18 input converter.

Sufficient module space was available in the IF processor to accommodate both input converters, the IF detector and switch S2. A second chassis contained the modulator, video detector and switch S1. Delayed AGC voltage was supplied to both input converters as explained under Figure 6.

In the actual case that inspired this example the switching was based on time and was controlled by a programmer. This doubtless is a more practical approach, but was modified for this example to show what could be accomplished using standard hardware.

### 2. Remote Demodulator Input Selection

A customer was picking up channels 2, 3, 6, 7, 8, and 38 at a microwave pickup point, and transmitting any two channels on a two-channel microwave link. Any of the six channels had to be supplied to either microwave transmitter. Remote selection of the channels is required. This is

accomplished using a tone signalling system which offers multiple contact closures upon receipt of unique tones from a transmitter.

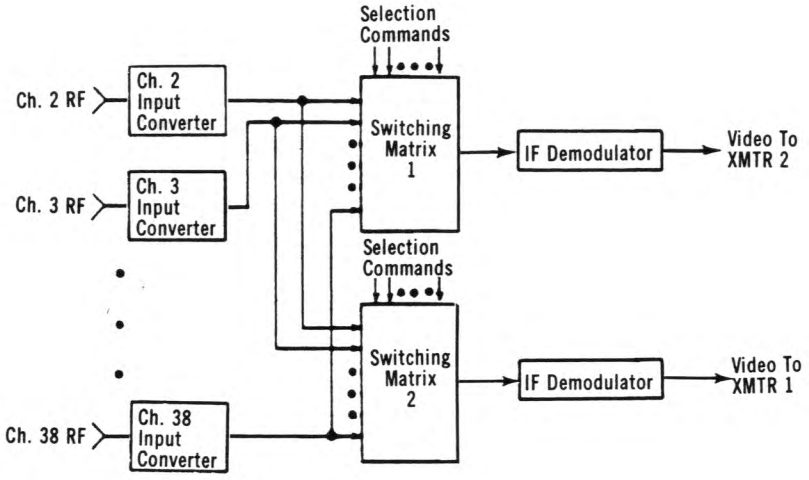



**Figure 11. Multiple Input Switching**

Six input converters are used as shown in Figure 11. The intermediate frequency output from each is split two ways and is applied to two identical switching matrices. Contact closures representing the selected channel are also supplied to the switching matrix. An IF demodulator receives as an input the selected channel, and supplies baseband video with a sound subcarrier to the microwave transmitter.

Each switching matrix was designed as a tree configuration and the small amount of decode logic required was added in discreet form. A rectangular matrix could have been employed to eliminate the need for logic, but the greatest possible isolation was achieved with the tree. (Realistically, since the solid state switches involved offered some 75 dB or more isolation, a rectangular matrix would have worked.)

For the particular case involved the customer had assured himself that the signal into the input converters would never exceed that at which delayed AGC action was required. Thus, all input converter delayed AGC inputs were tied to a voltage representing the maximum gain of the converter. Had delayed AGC been required, then switching would have been used to transfer it from the demodulators to the appropriate input converter.

All six input converters were housed in a modified processor chassis with a power supply. The splitters and solid state switches were mounted on an aluminum plate and could receive power from either the demodulator or input converter power supplies.

### 3. Multiple Video Switching

In this application the customer wished to supply 12 modulators from a pool of 29 video sources, each of which was accompanied by a 4.5 MHz sound subcarrier. If none of the 29 video sources is being supplied to a particular modulator, then it receives input from a character generator uniquely assigned to that modulator. At this time, the modulator was to accept a baseband audio signal.

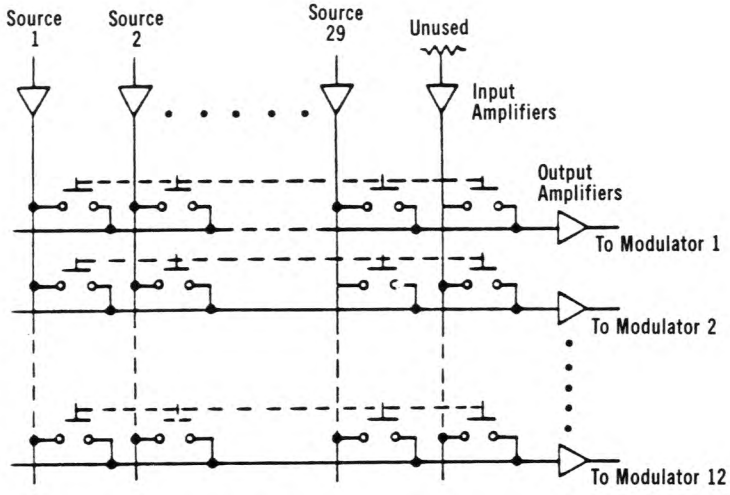



**Figure 12a. Routing Switcher Concept**

Figure 12a represents the concept of a routing switcher used to connect any of the 29 sources to any of the 12 modulators. Inputs are applied through buffer amplifiers, to a switch matrix. Only one switch in a row may be closed at a given time. This is generally accomplished by mechanically interconnecting pushbuttons in a row so that when one is depressed the previously activated button is released. Car radio pushbuttons work in a similar manner. The signal selected is supplied to the output amplifier at the end of the row. This output amplifier exhibits a high impedance input so that if more than one output is connected to the same input, amplitude shifting doesn't occur.

Use of a routing switch solves the problem of connecting any of the 29 inputs to any of the 12 outputs. Since all inputs to the routing switcher have a 4.5 MHz sound subcarrier included, then audio-follow-video is automatically assured. However, if none of these 29 inputs is selected, the modulator must accept input from a character generator and baseband audio. This function is accomplished as shown in Figure 12b.

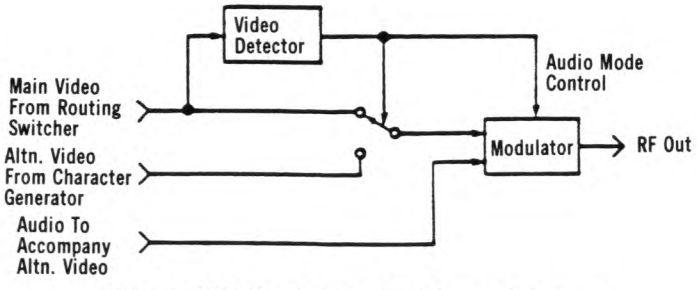



**Figure 12b. Switching to Alternate Source**

When none of the 29 inputs from the pool are selected, then column 30 for that row is selected on the routing switcher. This results in no input being supplied to the main video input on the modulator. The video switch then selects video from the character generator, and places the audio module in the mode to accept baseband audio rather than the subcarrier. This effectively switches audio sources without the use of an audio switch.

Had the same character generator been supplied to all sources, then it could have been supplied to the last input on the routing switcher. An auxiliary contact on the column 30 switch could have been used to change the sound mode of the modulator, or the sound could have been modulated onto a subcarrier before being applied to the switcher. However, with a separate character generator for each of 12 modulators, then the switcher would have to accommodate at least 41 inputs! Since the switcher cost goes up with additional inputs and since many cross-points would not be used, this would result in a very expensive switcher, with considerable wasted capability.

### 4. Multiple Fill IF Sources

In this case six programs were normally handled on six output channels. Six other programs were available which were to be used as fill as required on any channel. All switching was at IF, and remote control was provided through a dedicated telephone line. Tone signalling as described above was used to control switching.

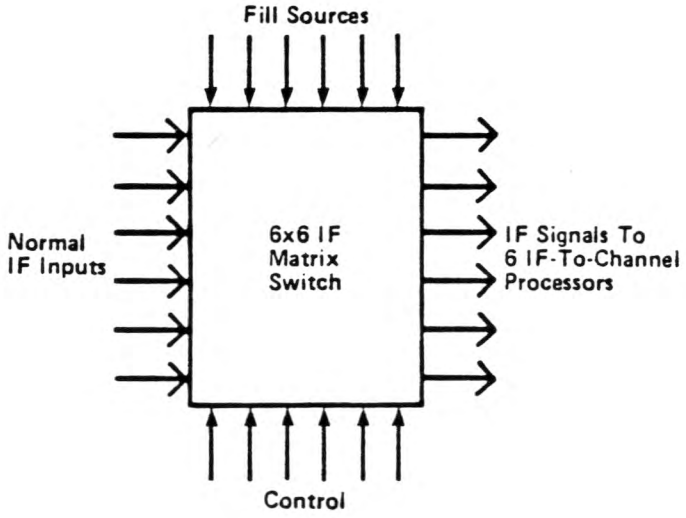



**Figure 13. IF Switching Block**

Figure 13 shows the basic switch matrix. One can see the similarities between this concept and the baseband routing switcher concept. The normal IF inputs were provided by breaking the IF path in a processor just before the output converter, as described in the earlier section on high level IF switching (Figure 5). IF signals from the switching matrix were returned to the originating processor. Fill IF signals were derived from a supertrunk originating at another site and pro-

cessed by channel-to-IF processors. Because these outputs had to be about 10 dB higher in level than normal, to accommodate splitting loss shown below, IF amplifiers were substituted for output converters.

RF cabling would have been simplified somewhat had the high level switch been placed in the processor and switched to a simplified fill source matrix. However, this would have complicated control wiring and the sequence of control operations that had to be programmed, so the chosen configuration resulted in a more "human engineered" approach.

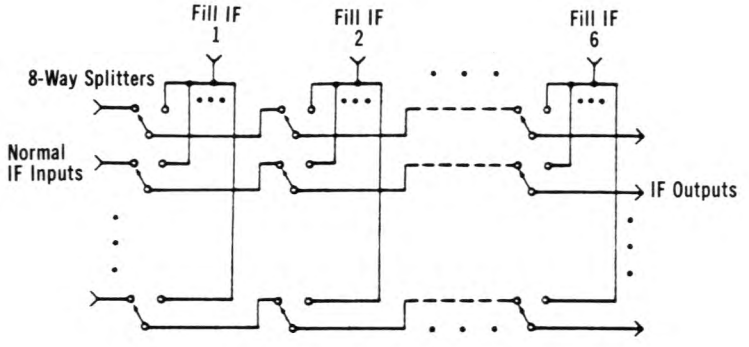



**Figure 14. IF Switching Matrix**

As shown in Figure 14, the splitting of the fill IF signals is accomplished using 8-way splitters. Two of the outputs are unused. In this case the tree distribution is chosen over a rectangular distribution which wouldn't need to have two unused outputs, because it was important to have the same signal level at all switches. Also, since all switches were in close proximity to each other, cabling simplification as a result of a rectangular distribution would not have been realized.

The switching configuration is rectangular as described above (for brevity, control signals are not shown). It was shown earlier that the rectangular configuration is potentially more susceptible to leakage from other sources than is a tree configuration. The author was not personally involved with checkout of this system, but calculates the total power of the six fill IF signals to be -62 dB from the main IF signal if the solid state switches used just met their 70 dB isolation specification. A typical isolation is around 75 dB, in which case the total power from the six fill sources would be -67 dB from the main IF. If a tree distribution pattern had been used, then depending upon the exact configuration, the total undesired power could have been down by so close to the switch isolation that the author's calculator barely was able to detect the difference (69.99999869 dB vs 70 dB). Thus, the good isolation of the switches used permitted a human engineered matrix that might not have been possible otherwise.

In practice, the isolation achieved may we.. have been more a function of leakage between cables than of switch isolation. To reduce cable leakage, high quality coaxial cable with two heavy shields was used. Connectors used had long crimp rings, with all crimped connections carefully

made using the manufacturer's recommended crimping tool. With these precautions, switching marices even more complex than this have been successfully installed. We can't emphasize too much, however, the need to use quality switching with excellent isolation, cable with heavy shielding, and carefully made connections.

## Conclusion

When contemplating a switching requirement, the designer should be aware of the trade-offs in switching hardware and techniques of implementation. Each design should take into consideration the types of signals to be handled and the ease with which the operator can comprehend his duties. Isolation is a major consideration in the design and implementation of switching. Factor affecting isolation include the isolation of switches used, cable leakage, cable routing, and integrity of shield ground connections. With proper attention to detail, either baseband or IF switching may be accomplished successfully.

Many options for switching hardware exist. As a general rule, IF switching hardware would be most likely obtained from CATV equipment manufacturers. Baseband switching equipment would probably be found in catalogs of broadcast equipment manufacturers, though CATV manufacturers are adding certain pieces of equipment to their product lines. Familiarity with the inner workings of equipment is essential if one is to reap the greatest benefit from dollars invested in switching hardware. Consulting manufacturers for advice on implementing a switching requirement will almost certainly bring useful advice on application of that manufacturers equipment.

All of the techniques discussed are oriented toward switching non synchronized signals: that is, signals which are not developed using a common sync source. Such switching is inevitably "dirty," permitting picture rolls and sometimes audible noise as well. Common broadcast practice dictates synchronous switching whenever possible. By operating all sources from a common sync generator and using a switcher that activates only in the vertical interval, much improved switching results. This is required for clean, roll-free switching, and should be the only switching considered in studio designs. However, the normal CATV switching applications involve video from varied sources which cannot be mutually synchronized, so one must expect "dirty" switching. Fortunately, the switching contemplated here occurs infrequently, and only between programs.

Finally, one feels compelled to ask whether or not a certain switching operation is really necessary. Granted, in many cases it is in order to get the greatest efficiency out of a channel or to solve regulatory requirements. However, we occasionally wonder what the effect on viewers is, if a switching operation is implemented which shifts expected programming to another channel or pre-empts it altogether. Certainly all switching operations should be held to a minimum consistent with regulatory restrictions and subscriber desires.

## Acknowledgements

The author wishes to express gratitude to Scientific-Atlanta application engineers Bob Mauney and Jerry Henning, who designed several of the switching facilities reported here. Their frequent

feedback to the design engineering group concerning customer applications is of great help in defining new requirements. Additionally, the author wishes to thank Alex Best, CATV division engineering manager at Scientific-Atlanta. Alex has contributed greatly to the solution of switching requirements through designs that reflect sensitivity to industry requirements.

## APPENDIX 1

### Calculation of Total Undesired Power from Multiple Sources

At several points in the text, the statement was made that with several switches each having a certain isolation, the total undesired carrier power was down so many decibels. This computation was made assuming equal power applied to all inputs of all switches. From the isolation of a switch, we can calculate the reduction in power through the unused port. For example, suppose that the isolation specification of a switch was 70 dB. This means that a signal applied to a non-selected input, will appear on the output 70 dB lower, or with $10^{-7}$ of the power it had at the input. If two switches are used in a simple rectangular matrix, Figure 15a, then the output power due to input B is $10^{-7}$ of the input power. The output power from input C is also $10^{-7}$ times the input to C. Thus, the total undesired output power is $2 \times 10^{-7}$ of the desired output power (assuming all equal input levels and no switch insertion loss). Expressed in decibels, $2 \times 10^{-7}$ represents -66.99 dB.

**Figure 15a**

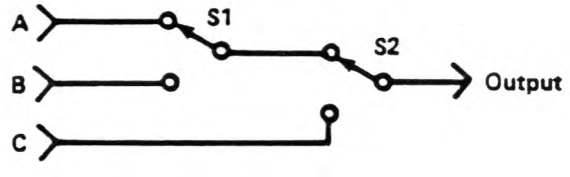



**Figure 15b**

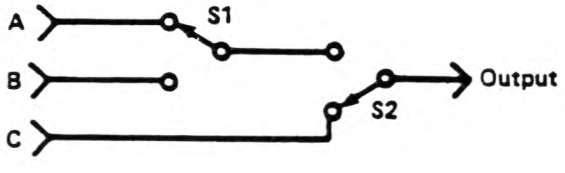



**Figure 15. Switch Isolation**

In Figure 15b switch S2 has been thrown to select input C. Through S1, input A is connected directly to the unused input of S2. Thus, the signal on A will appear on the output reduced by the 70 dB isolation of the switch, or at a power $10^{-7}$ of the input power. On the other hand, the signal on input B is reduced by a factor of $10^{-7}$ to the output of S1, and this power is again reduced by $10^{-7}$ at S2. Thus, the total output power contributed by B is the square of the isolation, or $10^{-14}$.

The total power contributed by A and B is then $10^{-7}$ plus $10^{-14}$ of the output power (again assuming all inputs at equal level), or -69.9999956 dB from the desired signal.

Note that we assume different inputs at slightly offset frequencies. The resultant interference in the output then has the form of a number of carriers of slightly different frequencies. We have been stating the relative power of the sum of all such interfering carriers. This should not be taken to have the same effect on a picture as one carrier with the same relative power.

## APPENDIX 2

### Ground Loops

Any time two pieces of equipment are interconnected, whether with switching or directly, the probability is that a ground loop will develop. Ground loops introduce noise, frequently but not necessarily from power mains, into a signal. The problem is generally more severe with video frequency signals, but may develop in RF circuits too. Ground loops take on a number of forms,

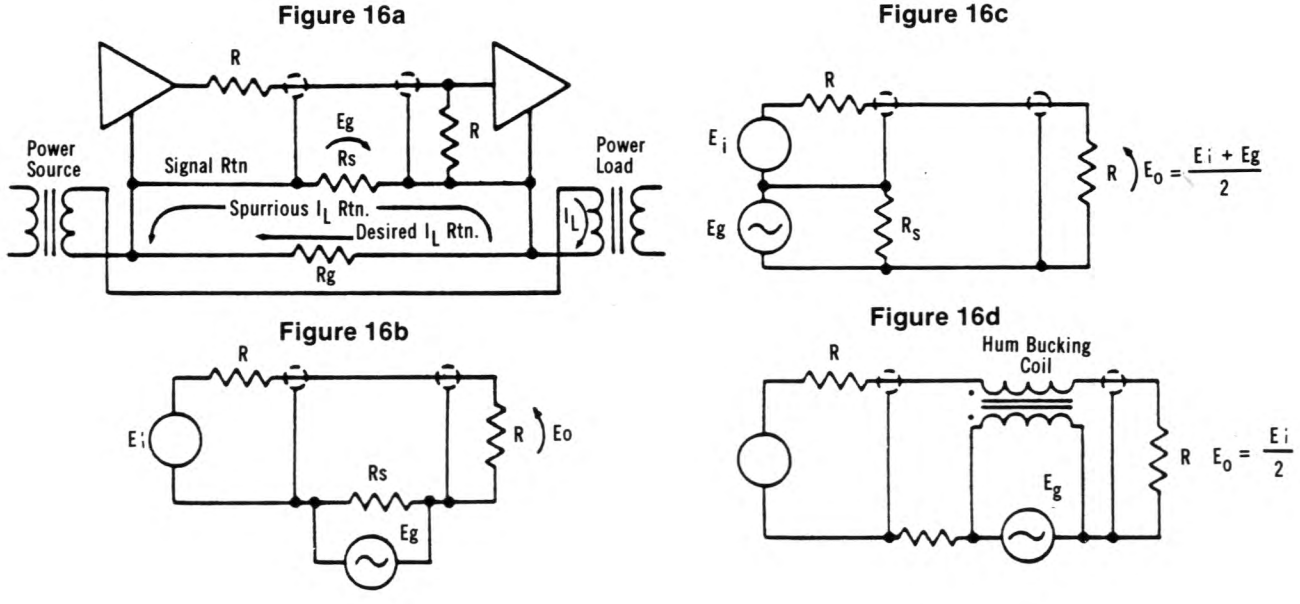



**Figure 16. One Source of Ground Loop Crosstalk**

one of which is shown in Figure 16. In A is shown a simple interconnection of two amplifiers located in different chassis and interconnected with coaxial cable. Also shown is a power circuit in which the power return is common with the shield return on each end. Return current from the power load should only flow in the power return, but since the power return is resistive, Rg, some current from the power load will return over the signal return path. This is shown in Figure 16B, where the power load return has produced a voltage $E_g$ on the signal return, which also has some

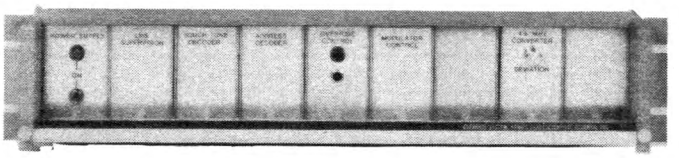
resistance. Figure 16B is redrawn in Figure 16C to show that this voltage in the ground circuit is in series with the source voltage $E_i$, and is thus transferred to the load. If the load doesn't include filtering which eliminates the voltage due to the ground loop, hum is superimiposed on the signal. One way of reducing the effect of the ground loop is to raise the common impedance of the center conductor and shield of the signal circuit. In Figure 16D is shown a hum bucking coil, a device which offers a high reactance to the ground loop current while not affecting the signal impedance between the coaxial center conductor and shield. The hum backing coil is made by winding small diameter coaxial cable on a ferrite core, producing an inductor when measured from one end of the winding to the other, but not affecting signal transmission through the coaxial cable.

Other methods for breaking ground loops are also frequently applied. One technique would be to reduce the impedance $R_g$, of the ground return. This may sometimes be accomplished by operating source and load equipment from a common voltage source. The author has sometimes succeeded in breaking a ground loop in the laboratory by simply plugging all test equipment into the same outlet strip.

Balanced signal transmission has long been popular as a way of circumventing ground loop and direct pick-up problems. Audio signal paths should always be designed using balanced signal transmission. For the maximum protection against ground loop problems, at least one end of the signal path is served with a transformer completely isolated from ground on the interface side. However, in many instances balanced transmission using non-transformer isolated lines is quite satisfactory. Video is sometimes transmitted on balanced lines when isolation from ground loops proves very difficult. Ground loops involving RF signals are frequently broken using transformers. In some cases loops have been broken by using two baluns of the type installed in the home, back-to-back. However, shielding integrity is compromised by the twin-lead on one side of the baluns, and by incomplete shielding of the balun housing.

After all the talk about breaking ground loops, we must finally admit, however, that eliminating ground loops can be a frustrating procedure involving more art than science. What works in one stance may not work in another. During design of an installation, one can carefully plan to isolate signal paths from power paths, to use balanced transmission of audio and high quality video circuits. After installation, though, experimentation may be required to eliminate ground loops.

## EMERGENCY AUDIO SYSTEM, MODEL 6120



**Features**

- All Solid State
- Fully Modular
- Access Code Guards Against Unauthorized Use
- Touch-Tone ® Telephone or Local Push-to-Talk Operation
- Processor or Modulator Audio Substitution

A convenient, secure method of distributing emergency messages simultaneously over all channels of this system is now available to the CATV operator with the Scientific-Atlanta Model 6120 CommAlert. Tornado, flood, and other severe weather warnings, as well as Civil Defense alerts can be broadcast over the cable system by authorized community officials. Access can be through the commercial telephone network or by a microphone located at the headend. Security against unauthorized use by telephone is afforded by a special access code programmed by the cable operator.

In some cases, the addition of this public service feature to a system can be funded through Civil Defense or could be justification for a rate increase.

Two modes of telephone control are provided. If a private, leased line is available, dialing the access code will interrupt program audio on all channels and place the telephone instrument on line. Audio is returned to normal when the telephone is hung up. If a commercial line is used, a telephone coupler, required for interconnection, will automatically answer the ringing line and cause the CommAlert to transmit dial tone to the calling party. The access code is then dialed and audio is placed on all channels from the telephone. Normal program audio will resume after a short period of silence.

If desired, the access-code feature can be omitted, allowing simple push-to-talk operation with a microphone.

For complete, flexible control of the headend audio, two types of outputs are provided in addition to a switching control line, allowing substitution of program audio by emergency audio via the headend signal processors or modulators. A 4.5-MHz audio subcarrier is provided for furnishing audio to signal processors. Three 600-ohm audio outputs are provided to furnish audio to modulators. Scientific-Atlanta Model 6150 Signal Processors will accept the 4.5-MHz audio subcarrier directly. Model 6350 Television Modulators equipped with Option -AS accept the 600-ohm audio line. Because of the high input impedance of the Model 6350 audio-switching option, many modulators can be driven from one CommAlert audio output.

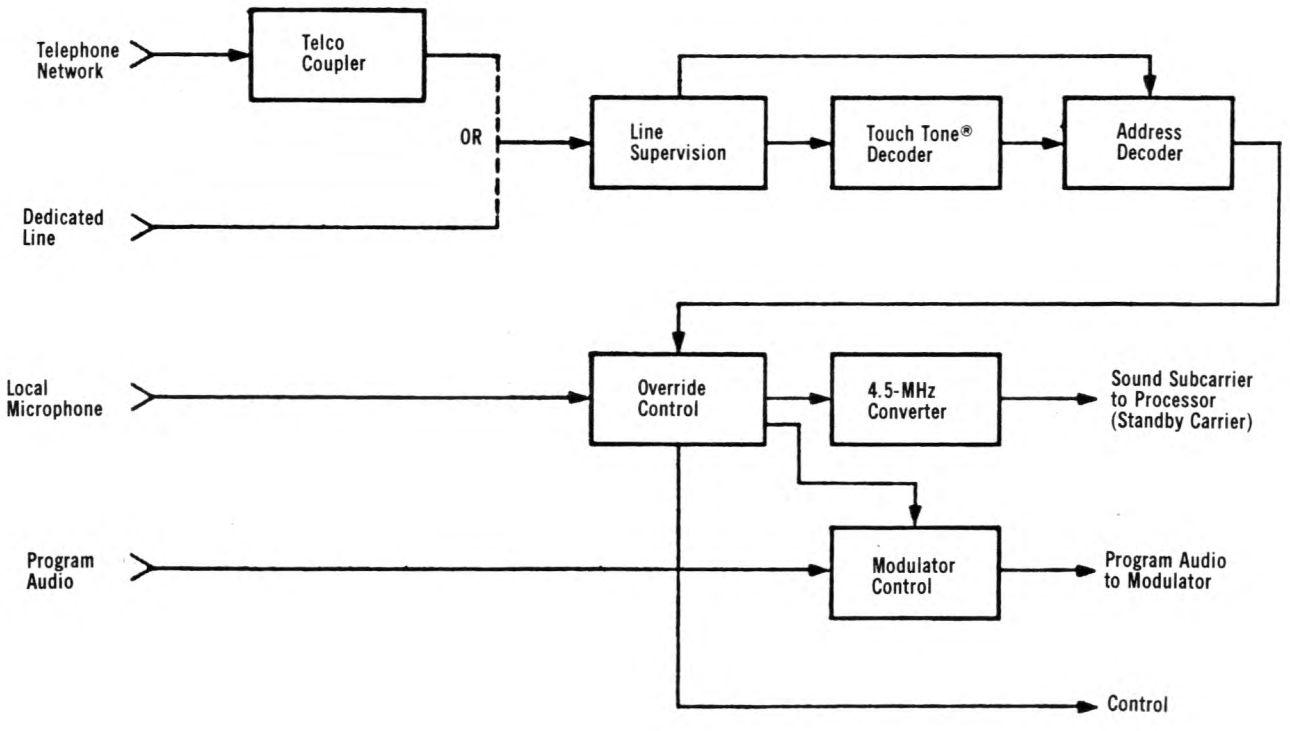



**Figure 17. Model 6120 CommAlert Block Diagram and Configuration for Audio Use Only**

## USING THE 6120 COMMALERT FOR BOTH VIDEO AND AUDIO EMERGENCY SWITCHING

If desired, both video and audio emergency switching may be accomplished with the 6120 CommAlert, 6350 Modulator and HIFS (High Level IF Switching) option. The CommAlert is addressed in the usual manner with a dedicated line or through the telephone coupler. The modulator accepts its audio input from the CommAlert once it has been activated. Video is also applied to the modulator where it is modulated to the intermediate frequency. Once modulated, the video and audio IF are amplified to a level of 60 dBmV and routed to the HIFS option in all the processors and modulators. The HIFS option is activated by a contact closure provided by the CommAlert. The nominal input to the HIFS switch is 37 dBmV. Exceeding this input level may result in spurious frequencies. Figure 18 is a block diagram showing the equipment required and order of connection.

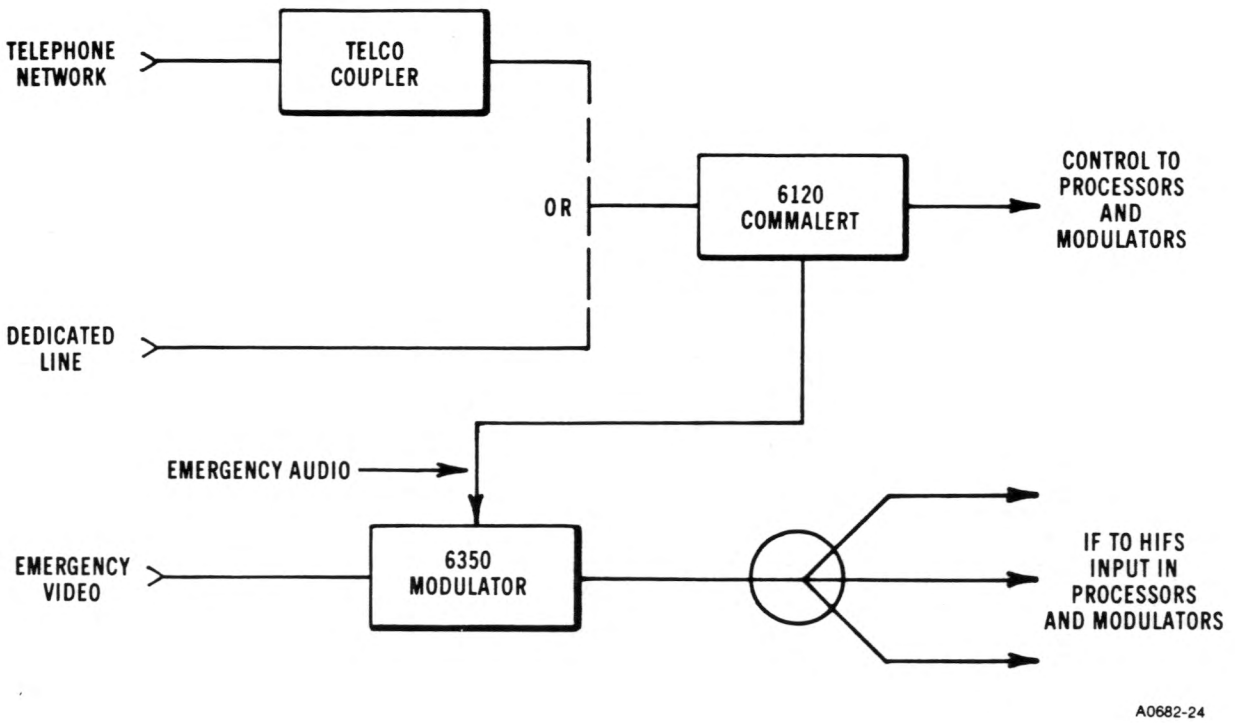



**Figure 18. Model 6120 CommAlert Configuration Diagram for Combined Video and Audio Use**

**Solid State Switch**

Used as the basic switching element in the 6150, 6250, and 6350 IF switching options, this device is now available as an accessory for use in headend switching systems. Housed in a small aluminum case, the switch can be mounted in any position. Operating as an SPDT coaxial switch, the device terminates the unused ports in 75 ohms. Switching is initiated by a closure to ground or by a positive voltage greater than +5V dc.

**Specifications**

Frequency Range
40 - 100 MHz

Maximum Level Input
+50 dBmV

Insertion Loss
<1 dB

Isolation
>70 dB at 50 MHz

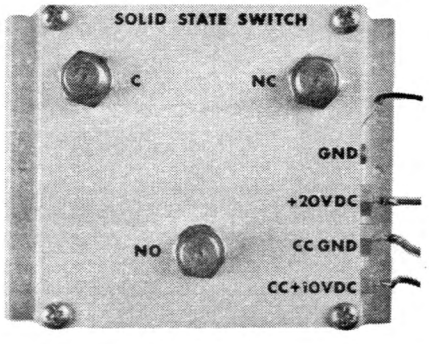
VSWR
<1.25:1

Transfer Time
0.1 sec. max.

Power Consumption
<500 mW

Control
Closure to ground
or +5V to +10V dc

Power Requirements
+20V dc, 25 mA max

Dimensions
102 mm wide (4 in.), 80 mm high (3.15 in.),
15.2 mm deep (0.6 in.)

Weight
Net: 0.18 kg (0.4 lb)
Shipping: 0.25 kg (0.56 lb)